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A DOPPLER RADAR CONTROLLED TRAFFIC INDICATOR

by

DENNIS L. CARR

A thesis submitted
in partial fulfillment of the requirements for the
Degree Master of Science, Department of

Electrical Engineering
South Dakota State University

1979

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A DOPPLER RADAR CONTROLLED TRAFFIC INDICATOR

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Dennis L. Carr

This thesis is approved as a creditable and independent investigation by a candidate for the degree, Master of Science, and is acceptable for meeting the thesis requirements for this degree. Acceptance of this thesis does not imply that the conclusions reached by the candidate are necessarily the conclusions of the major department.

Virgil G. Ellerbruch, Advisor
and Head, Electrical Engineering
Department

Date

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Dennis L. Carr

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CHAPTER I

INTRODUCTION

I-1. Problem Definition.

Traffic hazards on our highways are decreasing with the onset of new safety regulations, lower speed limits and highway design procedures. Unfortunately, economics prevent their complete elimination. New roadways are made safer while old roadways are pretty much ignored. Obstructions such as trees which restrict a motorist's view can be removed but it could be a monumental task to relocate other obstructions such as buildings, or to get the public to agree to the expense of leveling hills which lie too close to crossroads or private drives.

Should there be an obstruction such as a hill close to a driveway, a driver of the vehicle leaving the driveway to enter a main road has to guess whether or not there is a vehicle behind it. At night automobile headlights might provide the necessary warning as they are observed to shine onto nearby overhead lines or the tops of nearby trees. During the day the only warning would be the sound of the approaching vehicle, which in some cases would be muffled by other sounds or closed windows. A system is needed that would detect the presence of a vehicle behind such obstacles and warn motorists. The warning device could be placed alongside the roadway or fence line, depending on state and federal laws, with an indicator placed for easy visibility to motorists. Its design would have to comply with Federal laws and its operation would require the appropriate license. It would be desirable to house

the system so in such a way as to comply with present day ecological thought.

The system should be located in an enclosure about the size of a standard rural mailbox so that it could be easily set up alongside a roadway and not be obtrusive. Figure 1-1 is a pictorial view of the problem and the proposed solution.

Figure 1-2 shows the block diagram of a system that could be used. This system processes the doppler frequency created by mixing the transmitted signal with the signal reflected from an approaching vehicle and then turns on a warning indicator. The indicator is turned on if vehicle velocity is within the range from 30 miles per hour (mph) to 80 mph. Vehicles exceeding 80 mph will be detected but not as soon as those vehicles with speeds between 30 mph and 80 mph. The reason for this is that the radar system detects only the radial velocity between itself and the vehicle. As the radar system is to the side of the path of the vehicle the radial velocity will vary from the speed of the vehicle, when it is at a distance, to zero as it passes the radar set. This will be explained in more detail in a later chapter.

Six seconds is sufficient time for the entering vehicle of Figure 1-1 to either make a left hand turn and enter the appropriate lane of the main roadway or make a right hand turn and gain some speed. If the approaching vehicle is traveling 60 mph (5 mph over the speed limit) it will travel 88 feet per second. This will require that it be

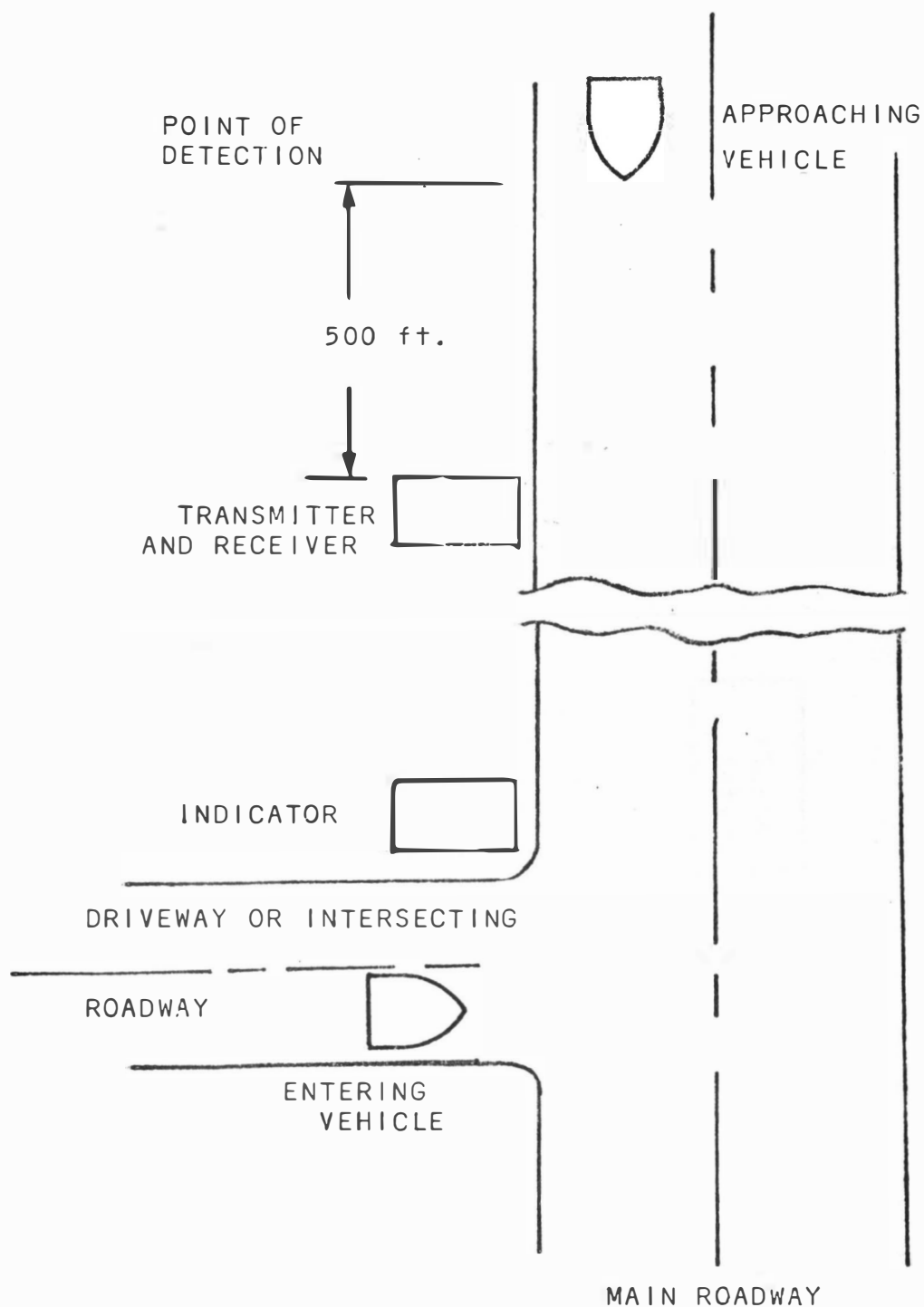


FIGURE 1-1. TRAFFIC HAZARD LAYOUT.

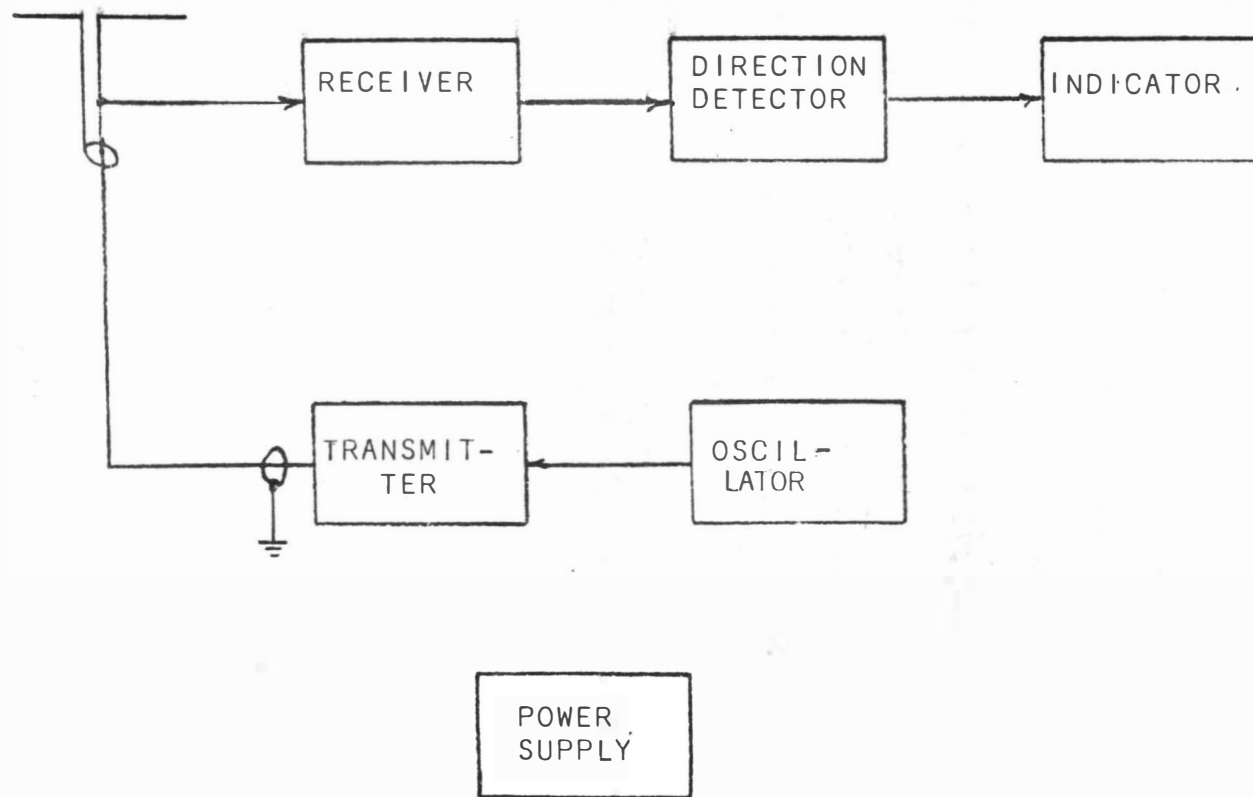


FIGURE 1-2. BLOCK DIAGRAM OF DOPPLER SWITCH.

detected about 500 feet from the transmitter. The distance between the transmitter and indicator will provide an additional margin of safety.

Most radar sets used for detection operate in the optical range, where the signal wavelength is much smaller than the object's width. The proposed system will operate in the Mie region, where the signal wavelength is approximately equal to the target's width. This will result in a somewhat weaker reflected signal than in the optical range, but for a range of 500 feet the signal received by the receiver will have sufficient power to be detected. This is shown in Chapter III.

1-2. Specifications.

The average car is approximately six feet wide. The wavelength (λ) for 500 mHz, is:

$$\lambda = \frac{C}{F} = \frac{3 \times 10^8}{5 \times 10^8} \text{ meters} = .6 \text{ meter}$$

$$\lambda = .6 \text{ meter} \times \frac{3.28 \text{ ft.}}{\text{meter}} = 1.97 \text{ feet.}$$

This can be considered as being in the Mie region. Other reasons for choosing a frequency of 500 mHz are as follows:

- The signal at 500 mHz will be virtually free from attenuation due to atmospheric and weather conditions. Atmospheric attenuation is about 0.006 db/km (1).
- The wavelength is long with respect to birds, therefore they will reflect very small portions of the incident wave's energy

and will have little, if any, effect on the system.

- The wavelength cannot be much longer than the target diameter or too much power will be lost (scattered), resulting in a reflected signal which is too weak to detect.
- A lower frequency will require an antenna too large for an enclosure the size of a standard rural mailbox.

The general specifications for the system are as follows:

Range = 500 feet.

Vehicle Velocity for Detection = 30 mph minimum, 80 mph maximum
at 500 feet from the transmitter.

Frequency = 500 MHz.

Transmitter Power = 0.1 watt.

CHAPTER II

BASICS OF RADAR DESIGN

2-1. Historical Development of Radar.

Radar, an acronym for radio detection and ranging, was adopted as the official name for the detection of objects by the radiation of electromagnetic waves. The development of radar can be broken into six development stages, or discoveries:

- (1) Radio waves are reflected by abrupt changes in dielectric constant, permeability and conductivity.
- (2) Radio waves can be directed along narrow beams.
- (3) Distance can be determined by timing the travel of a radio wave.
- (4) Reflected waves may be detected by wave interference patterns.
- (5) A pulsed radar can be used to measure distance.
- (6) The combining of the previous five stages into one unit (1).

The first electromagnetic wave detection of an object was accomplished by accident. In 1930, L. A. Hyland of the Naval Research Laboratory was working with a direction-finding device when he noticed that every time an aircraft flew through the electromagnetic beam the returning signal intensity increased. This discovery created an increased interest in radar and its development began.

The first radars, called CW or continuous wave radars because they transmitted a continuous or uninterrupted wave of energy, could only provide the position of an object in one dimension. A more useful radar

system will provide an object's position in at least two of the three dimensions height, azimuth and range.

A pulsed system was then developed which transmitted a short burst of energy and measured the round trip time to travel to the object and reflect back to the receiver. Using the measured time and the knowledge that the speed of radiated electromagnetic energy is approximately the speed of light, the distance to the object is:

$$\text{Distance} = \text{Rate} \times \text{Time}. \quad (2-1)$$

The pulse geometry is shown in Figure 2-1.

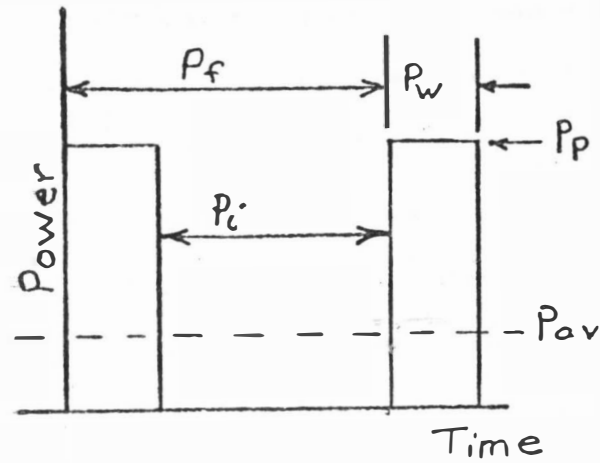
By scanning the antenna in a horizontal plane the azimuth of the target is determined. Azimuth is the angular distance the antenna has scanned from a reference direction to the direction the antenna is pointing the moment a pulse reflected from the object is received. By scanning the antenna in a vertical plane, the height of the target can be determined using the distance to the target and the angle between the antenna's bore sight and a horizontal reference.

$$\text{Height} = \text{Distance} \times \sin \theta \quad (2-2)$$

where

θ = Angle between the antenna's bore sight and a horizontal reference.

The first radars were operated in the frequency range from 20 MHz to 60 MHz. The British, though, developed a 12 MHz radar in 1935 which had a range of 40 miles. By 1939, the British, in a frantic effort to build up their defenses because of unrest in Europe, had developed a 200 MHz airborne radar system. While developing this radar system they discovered that as the radar's frequency was increased its resolution



P_w = Pulse Width

P_f = Pulse Period

P_i = Pulse Interval

P_{av} = Average Power

P_p = Peak Power

Duty Cycle = $\frac{P_w}{P_f}$

$P_{av} = P_p \frac{P_w}{P_f}$

FIGURE 2-1. PULSE GEOMETRY.

improved. Resolution is the ability to distinguish between two separated objects and is a function of frequency and beamwidth. Beamwidth is the width of the radiation pattern transmitted from the antenna and is a function of wavelength and antenna reflector geometry (1).

Herein lies the reason for microwave use. Higher frequencies provide better resolution. An additional advantage is that higher frequencies permit the use of smaller antennas and waveguides. Because microwave radar sets use smaller antennas a greater variety of antennas can be used. The microwave frequency bands used by radar systems today are listed in Table 2-1 (1).

2-2. The Doppler Frequency Shift.

The doppler phenomenon was named after the Austrian mathematician and physicist, Christian Doppler, who first predicted it. When electromagnetic energy is incident on an object it undergoes a frequency or doppler shift proportional to the relative velocity between the object and the energy source. This doppler frequency shift can be used for determining the radial velocity of an object moving with respect to the radar transmitter.

The doppler frequency, when the source and target are receding is:

(2) (3)

$$f_{\text{doppler}} = f_{T_X} \left[\frac{1 - V_r/c}{\sqrt{1 - (V_r/c)^2}} \right] \quad (2-3)$$

where:

f_{doppler} = shifted or doppler frequency

f_{T_X} = transmitted frequency

TABLE 2-1. MICROWAVE FREQUENCY BANDS.

MICROWAVE FREQUENCY BAND	FREQUENCY
UHF	300 - 1000 MHz
L	1000 - 2000 MHz
S	2000 - 4000 MHz
C	4000 - 8000 MHz
X	8000 - 12500 MHz
Ku	12.5 - 18 GHz
K	18 - 26.5 GHz
Ka	26.5 - 40 GHz
Millimeter	40 GHz and above

V_r = radial velocity of object with respect to the
transmitter

C = speed of light

For $V \ll C$

$$f_{\text{dopper}} \approx f_{T_X} \left[1 - V_r/C \right] \quad (2-4)$$

It must be remembered that equation (2-4) gives the frequency of the incident signal as detected by an object moving with radial velocity V_r . This is also the frequency of the signal reflected, transmitted passively, from the object. This signal is then shifted in frequency a second time when it is detected by the receiver. The received signal is mixed with part of the transmitted signal and the difference in frequency between the two signals is processed to determine the object's radial velocity. The equation for the difference between the received and transmitted signals is:

$$\begin{aligned} F_D &\approx f_{T_X} \left[1 - \frac{V_r}{C} \right] \left[1 - \frac{V_r}{C} \right] - f_{T_X} \\ &\approx f_{T_X} \left[1 - \frac{V_r}{C} \right]^2 - f_{T_X} \\ &\approx f_{T_X} \left[1 - \frac{2V_r}{C} + \frac{V_r^2}{C^2} \right] - f_{T_X} \end{aligned} \quad (2-5)$$

$$\begin{aligned} F_D &\approx f_{T_X} - 2f_{T_X} \frac{V_r}{C} + f_{T_X} \frac{V_r^2}{C^2} - f_{T_X} \\ F_D &\approx - \frac{2f_{T_X} V_r}{C} + f_{T_X} \frac{V_r^2}{C^2} \end{aligned}$$

When

$$\frac{V_r}{C} \ll 1$$

then

$$\frac{V_r}{C} \gg \frac{V_r^2}{C^2}$$

therefore

$$F_D \approx -2f_{TX} \frac{V_r}{C} .$$

Let

$$V_r = V \cos \theta$$

then

$$F_D \approx - \frac{2f_{TX} V \cos \theta}{C} \quad (2-6)$$

where

$$V_r = V \cos \theta$$

V = Velocity of the object

θ = Angle between the direction of the object's motion
and the direction to the transmitter.

F_D = Total change in frequency between transmitted
and received signals.

If the direction coincident with the direction of the transmitted signal is assigned positive then the minus sign in equation (2-6) indicates that F_D is negative for receding objects and positive for approaching objects.

2-3. Radar Design.

The beginning of a radar design begins with the determination of its specifications. A good place to start is by examining the basic radar equation (1)

$$P_r = \frac{P_t A_e^2 \sigma}{4\pi \lambda^2 R^4} \quad (2-7)$$

where R = Distance to object

P_t = Transmitted power in watts

$G = \frac{4\pi A_e}{\lambda^2}$ where A_e = Effective antenna area in square meters for a half wave dipole.

λ = Wavelength in meters

$$= \frac{3 (10)^8}{\text{Frequency}}$$

σ = Effective area of the object in square meters.

P_r = Power amplitude of signal at the receiver.

For $S_{\min} = P_r$

$R_{\max} = R$

$$G = G_t = G_r = \frac{4\pi A_e}{\lambda^2}$$

and the transmitting antenna is also the receiving antenna then

$$S_{\min} = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 R_{\max}^4} \quad (2-8)$$

where S_{\min} = Minimum detectable signal by the receiver in watts

P_r = Received power in watts

G_t = Gain of transmitting antenna

G_r = Gain of receiving antenna.

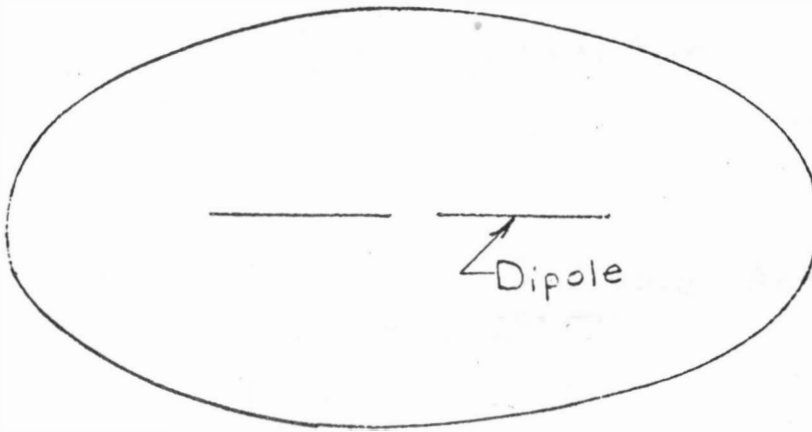
The maximum range of a radar set might be determined by its use. For example, there may be range restrictions imposed to eliminate interference with nearby radar sets, or in the case of the proposed system it will be impractical to detect vehicles much further than 150 meters from the transmitter. To do so would cause a motorist to wait at an intersection an unnecessary length of time.

The transmitted power can remain an unknown provided all the other variables are known. Where FCC regulations impose maximum power restrictions then that value could be used for P_t , while some other variable such as range is left as an unknown.

The effective area of an antenna is the cross-sectional area enclosing the antenna and having an amount of power equal to the power captured by the antenna. Figure 2-2 shows the effective area for a dipole antenna (4).

The power gain of a dipole antenna can be increased 6 db by using a reflector to make the antenna directional (4). A flat reflector one-quarter wavelength from a dipole antenna is shown in Figure 2-3. Figures 2-4 and 2-5 show two other types of antenna reflectors commonly used (2). The parabolic reflector is the most directional and therefore is the most common type of antenna used on radar sets.

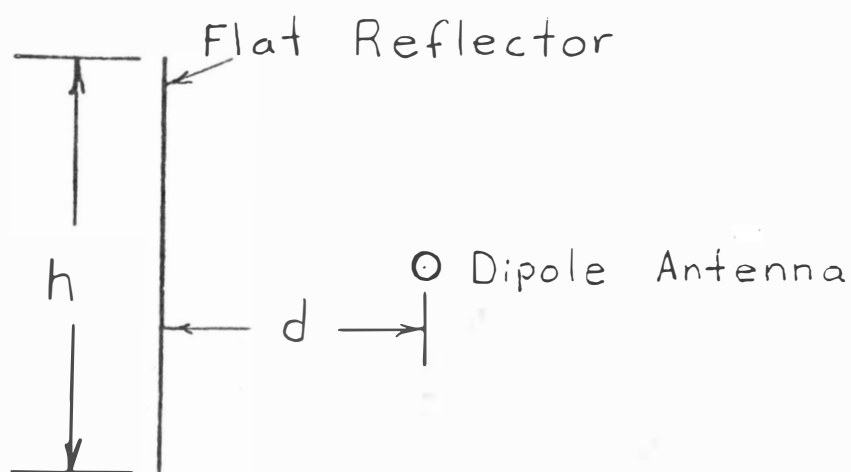
The geometry of the object to be detected is always known, therefore its effective area if not known can be empirically determined by solving



For a half wave dipole

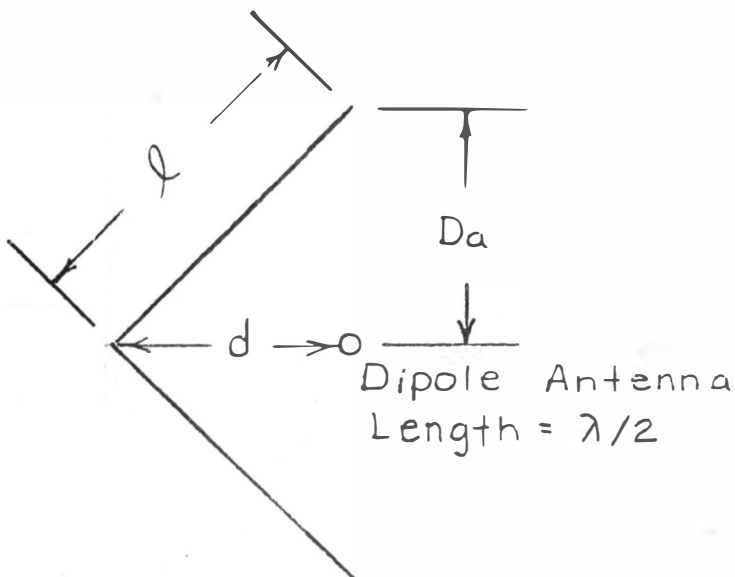
$$A_e = \frac{G \lambda^2}{4 \pi}$$
$$= 0.131 \lambda^2$$

FIGURE 2-2. EFFECTIVE AREA OF A DIPOLE ANTENNA.



$d = 0.25 \lambda$
 $A_v = 6 \text{ db}$
 $h = \text{Height of reflector}$

FIGURE 2-3. DIPOLE ANTENNA WITH FLAT REFLECTOR.



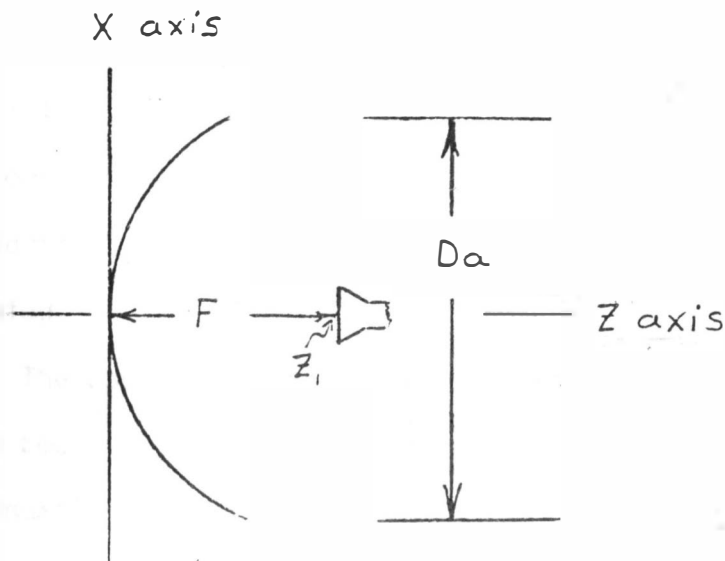
$$D_a \geq \lambda$$

$$d = 0.25 \lambda \text{ to } 0.7 \lambda$$

$$\text{Length of reflector} = 2d$$

$$\text{B.W.} = 40^\circ$$

FIGURE 2-4. CORNER REFLECTOR ANTENNA.



$$X^2 = 4 F Z_i$$

For $Z_i = \text{Focal Point}$

$$D_a = 4 F$$

$$B.W = 70 \lambda / D_a$$

For illumination tapered to
-10 db at edges.

FIGURE 2-5. PARABOLIC REFLECTOR ANTENNA.

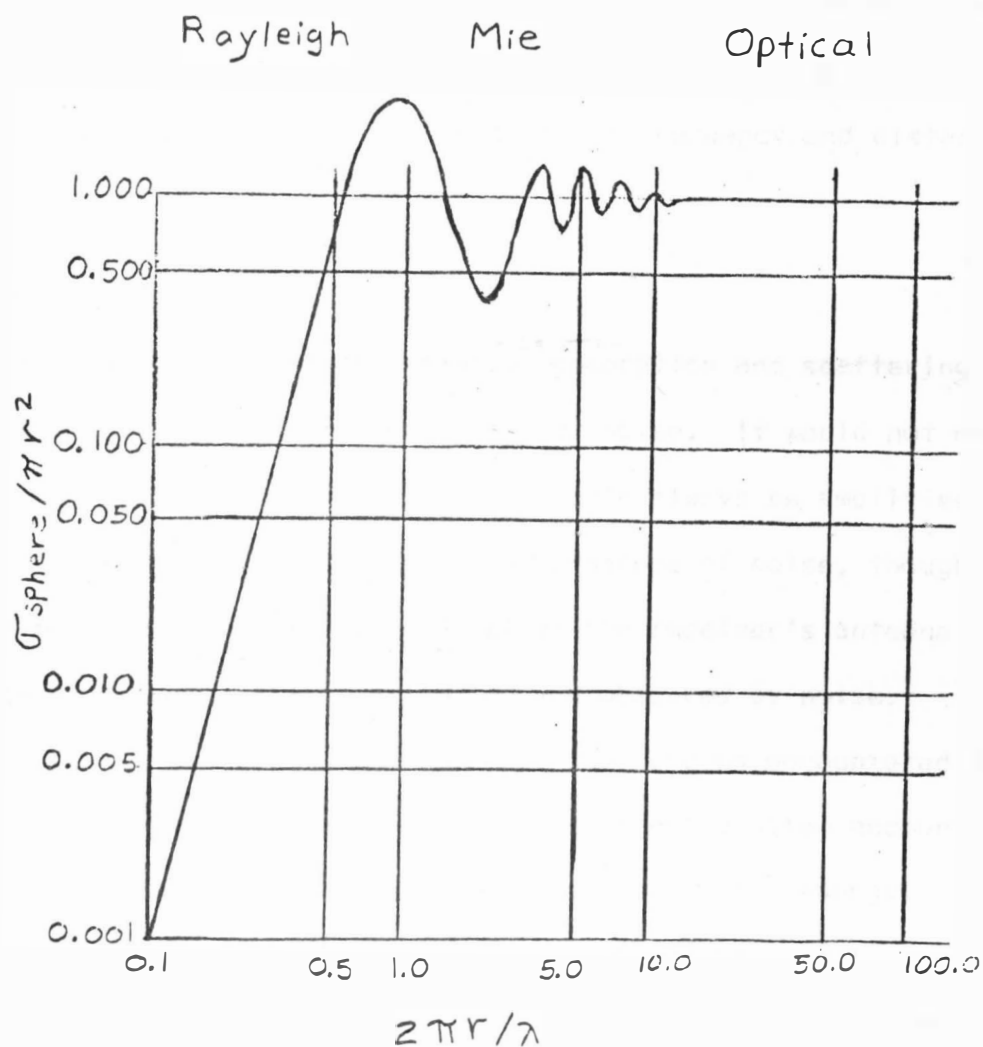
equation 2-3 for σ , and assigning values for the remaining variables except S_{\min} . Then a test situation is set up where a power magnitude equal to P_t and a wavelength, λ , is transmitted to the object over a known distance. The reflected signal is then measured and substituted for S_{\min} in the radar equation. A value for σ can then be calculated.

The basic radar equation will not produce exact results in practice because it does not include signal degradation due to atmospheric attenuation and noise (1).

2-4. Atmospheric Attenuation.

Figure 2-6 defines the types of scattering from a spherical object (1). Rayleigh scattering is signal attenuation due to scattering from a purely gaseous atmosphere. The Rayleigh region is where the signal wavelength is larger than the circumference of the object detected, assuming the object to be a sphere. Mie scattering is signal attenuation due to absorption and scattering by small dielectric spheres when the signal wavelength is approximately equal to the circumference of the object detected.

The optical region shown in Figure 2-6 is where the object circumference is much larger than the signal wavelength. Although some power absorption occurs in this region, most of the signal is reflected. A radar operating in the optical region will consistently receive the largest signals of all three regions for a larger variety of object sizes.



σ = Power reflected back to the receiver per unit solid angle
Incident power density / 4π

r = Radius of sphere

λ = Wavelength of incident signal

FIGURE 2-6. NORMALIZED RADAR CROSS SECTION OF A SPHERE VERSUS CIRCUMFERENCE IN WAVELENGTHS.

Attenuation due to absorption in the atmosphere is primarily due to water vapor, carbon dioxide and ozones, although oxygen and nitrogen absorption occurs for shorter wavelengths. Figure 2-7 and Figure 2-8 show atmospheric attenuation as a function of frequency and distance (1).

2-5. Noise Problems.

Signal attenuation due to distance, absorption and scattering would not present any problems if it were not for noise. It would not matter how much a signal had been attenuated it could always be amplified and restored to its original amplitude. The presence of noise, though, establishes a minimum useful signal level at the receiver's antenna. Signals below this level will be distorted and obscured by noise.

Noise can be defined as any unwanted frequencies encountered while processing a desired signal. Noise is the biggest problem encountered in the design of receivers for radiated electromagnetic energy.

A broad categorization of the types of noise are:

- 1) Noise generated in the transmitter and radiated along with the signal.
- 2) Noise from sources external to either the transmitter or the receiver.
- 3) Noise generated at the antenna.
- 4) Noise generated in the receiver.

Proper design techniques can be used to reduce the level of noise prior to transmission. Noise generated externally to the transmitter and the receiver is more difficult to eliminate. It can enter the receiver

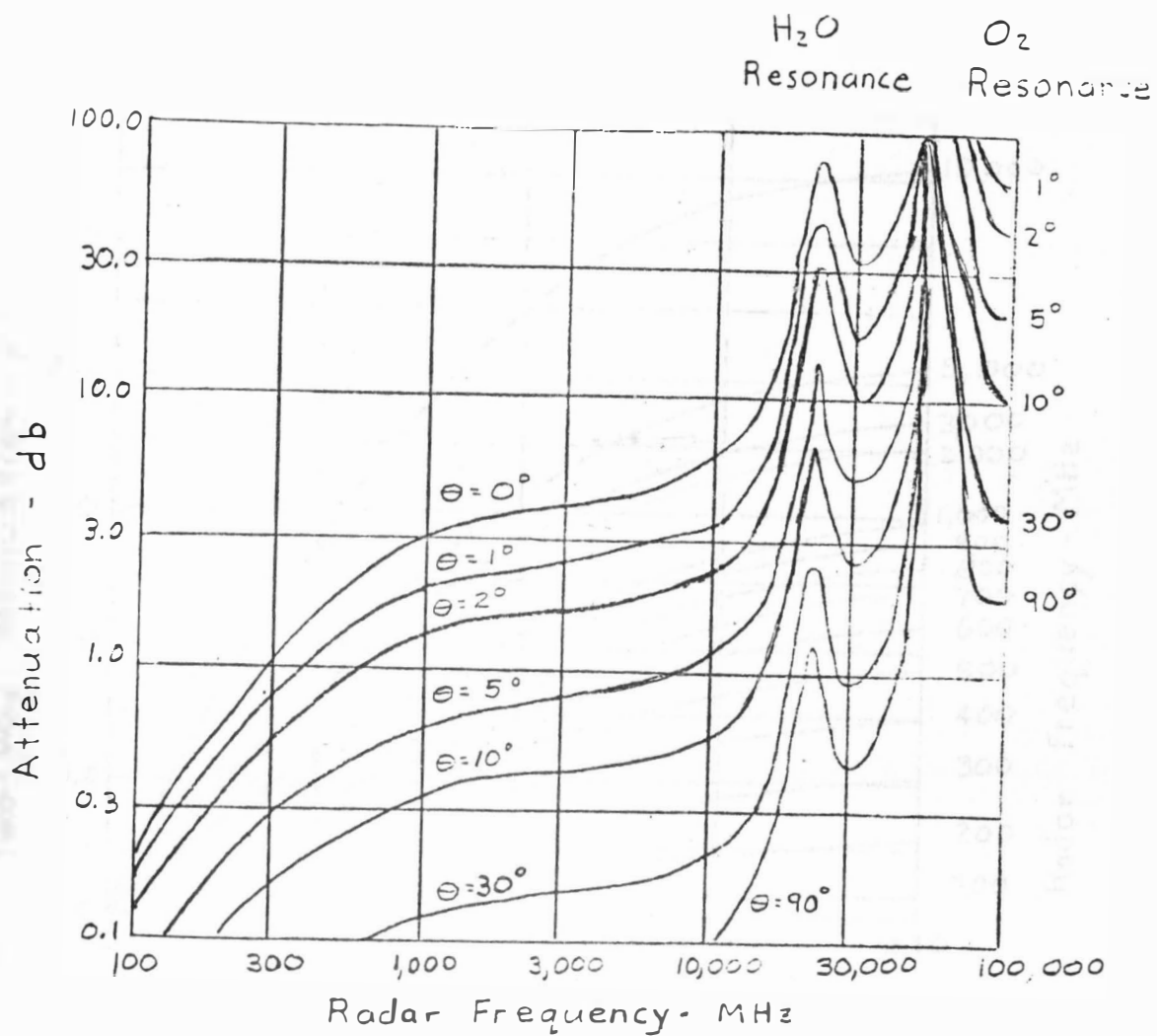


FIGURE 2-7. TWO-WAY RADAR ATTENUATION FOR ENTIRE TROPOSPHERE
VERSUS RADAR FREQUENCY AT VARIOUS ELEVATION ANGLES, θ .

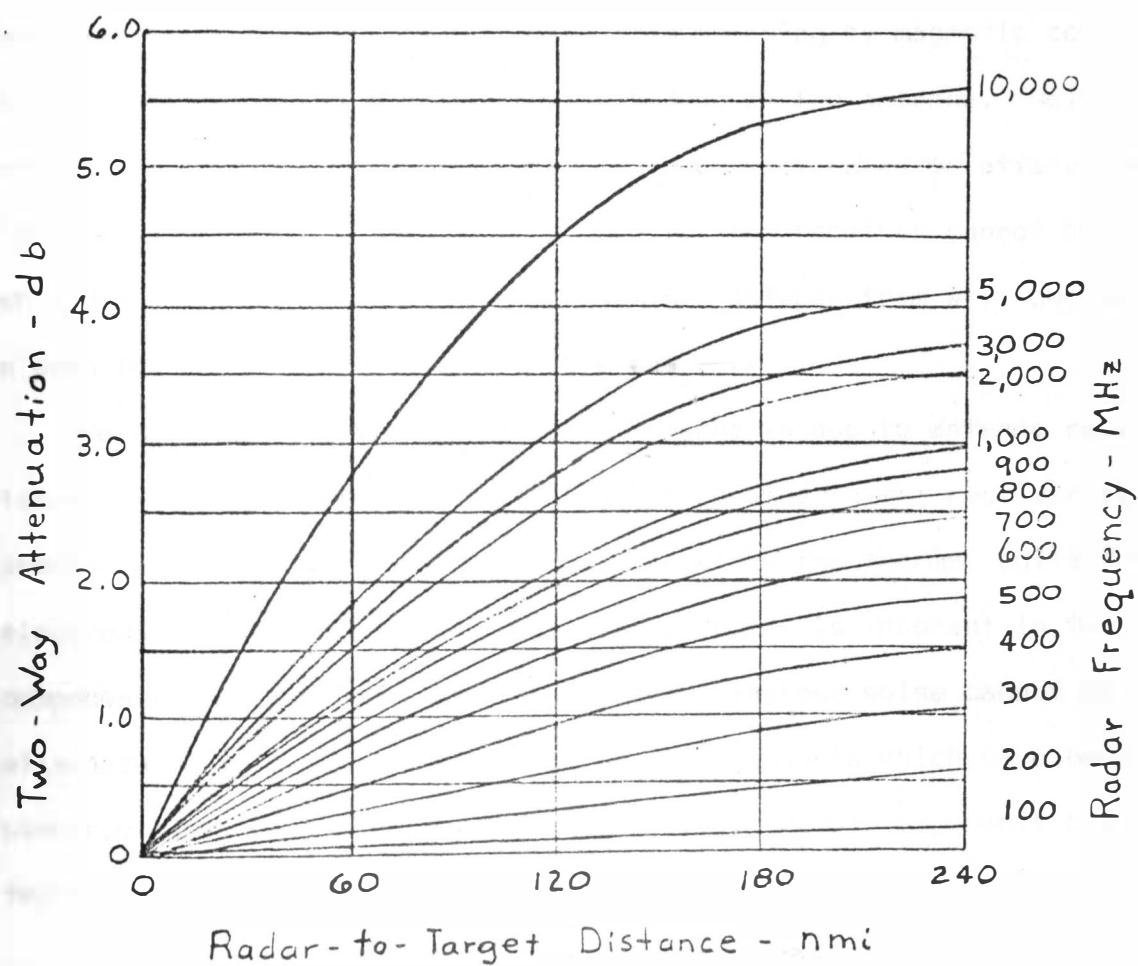


FIGURE 2-8. TWO-WAY RADAR ATTENUATION AT SEA LEVEL VERSUS DISTANCE AT ZERO DEGREES ELEVATION.

by either magnetic coupling or radiation. Shielding the receiver with a metallic material will prevent noise from entering by magnetic coupling and by that radiation which is not picked up by the antenna. Noise which enters the receiver by the antenna has to be eliminated or attenuated in some other way. If the power source for the receiver cannot be shielded with the receiver, the power lines between them will become a path for noise unless they are shielded..

Noise generated at the receiving antenna is due to antenna resistance and cannot be avoided (5). Noise generated in the receiver is predominantly thermal noise, which is caused by the thermal agitation of electrons in the components. This type of noise is inherent in the components used for receiver construction. Thermal noise cannot be eliminated but it can be reduced by using components which have been constructed to be low in thermal noise. Using these components and proper techniques in design, thermal noise can be kept at a minimum.

A comprehensive study of noise and its control is beyond the scope of this thesis. The reader, if interested is referred to references (5), (6), (7) and (8) for a more complete study of noise.

2-6. Complex Circuit Components.

At microwave frequencies standard components like capacitors, resistors, and inductors begin looking like combinations of each other. In other words they become complex. The reasons are because any length of wire has a resistance and an inductance associated with it regardless how small it is, and any dielectric with conductors on either side of it

is a capacitor. Therefore, a resistor has its normal value of resistance plus the inductance of its leads and the capacitance between the leads with the resistance material acting as a dielectric. An inductor has the resistance of its wire coils and each coil is separated from another coil by the dielectric material used for insulating the turns. A capacitor has the inductance of its leads and plates and the resistance of the wire and dielectric material. A component then takes on the value which dominates at a particular frequency. Figure 2-9 shows the equivalent models for a capacitor, a resistor and an inductor.

Analyzing the capacitor model shows why its impedance is a function of frequency.

$$Z_c = jX_L' + R + \left[\frac{-jX_c R'}{R' - jX_c} \right] \quad (2-9)$$

$$= R + j\omega L' + \frac{-jR' / \omega C}{R' + 1/j\omega C}$$

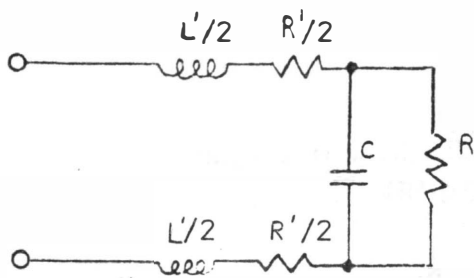
where

$$\omega = 2\pi F$$

$$F = \text{Frequency}$$

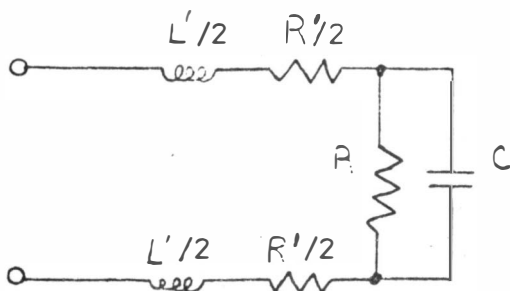
After some mathematical manipulation

$$Z_c = R + \frac{R'}{(R' \omega C)^2 + 1} + j\omega \left[L' - \frac{(R')^2 C}{(R' \omega C)^2 + 1} \right] \quad (2-10)$$



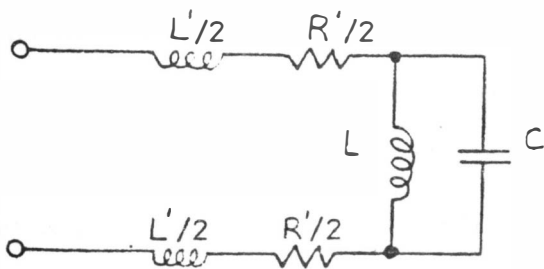
L' = Lead Inductance
 R' = Lead Resistance
 C = Capacitor
 R = Resistance of Dielectric

CAPACITOR MODEL



L' = Lead Inductance
 R' = Lead Resistance
 R = Resistance
 C = Capacitance between Leads

RESISTOR MODEL



L' = Lead Inductance
 R' = Lead Resistance and Coil Resistance
 L = Inductance
 C = Shunt Capacitance

INDUCTOR MODEL

FIGURE 2-9. CIRCUIT COMPONENT MODELS.

with

$$\text{Real} = R + \frac{R'}{(R'\omega C)^2 + 1}$$

$$\text{Imaginary} = j \left[\omega L' - \frac{\omega CR'^2}{(R'\omega C)^2 + 1} \right]$$

As $\omega \rightarrow 0$

$$Z_c \rightarrow R + R'.$$

As $\omega \rightarrow \infty$

$$Z_c \rightarrow \infty. \quad (2-11)$$

This seems correct because a capacitor presents a high impedance to dc and its impedance decreases as frequency increases. But this equation does not show R and R' to be frequency sensitive. Skin effect has not been considered. Briefly, skin effect is a name given to the phenomenon of conductor current crowding to the surface of the conductor as frequency increases. Obviously, if the current is contained in a smaller cross-sectional area it encounters a proportionally higher resistance. The effective resistance considering skin effect, is

$$R = \frac{L}{2\pi a \sigma \delta} \quad (2-12)$$

where

L = Wire length in meters

a = Wire radius

σ = conductivity = $5.88(10^7)$ /m for copper

δ = skin depth

$$= 1/\sqrt{\pi f \mu \sigma}$$

f = frequency

$\mu = \mu_r \mu_o$ = permeability

$\mu_o = 4\pi \times 10^{-7}$ h/m for copper

μ_r = relative permeability.

There are other losses which manifest themselves as impedance changes such as radiation, inductive and capacitive effects between components.

The most significant change in capacitors is due to power losses in the dielectric. For low loss dielectric capacitors with short leads, capacitance is almost frequency insensitive (9).

Similar equations for the resistor and inductor models can be derived. Impedance changes that affected the capacitor also apply to the resistor and the inductor, but for the resistor and inductor skin effect causes the greater impedance variations with frequency.

The equation for the resistor model is

$$Z_R = R' + \frac{R}{(\omega C R)^2 + 1} + j\omega \left[L' - \frac{C R}{(\omega C R)^2 + 1} \right] \quad (2-13)$$

Here, again, the equation is not accurate as lead inductance and the resistance has been changed due to skin effect; therefore L' , L and R must be modified accordingly.

The equation for the inductor model is

$$Z_L = R' + j \left[\omega L' - \frac{1}{\omega C - \frac{1}{\omega L}} \right] \quad (2-14)$$

As can be expected skin effect has affected this model, also. Inductance is affected because as the current at high frequencies is forced to the surface of a conductor the internal inductance decreases as can be seen from (9)

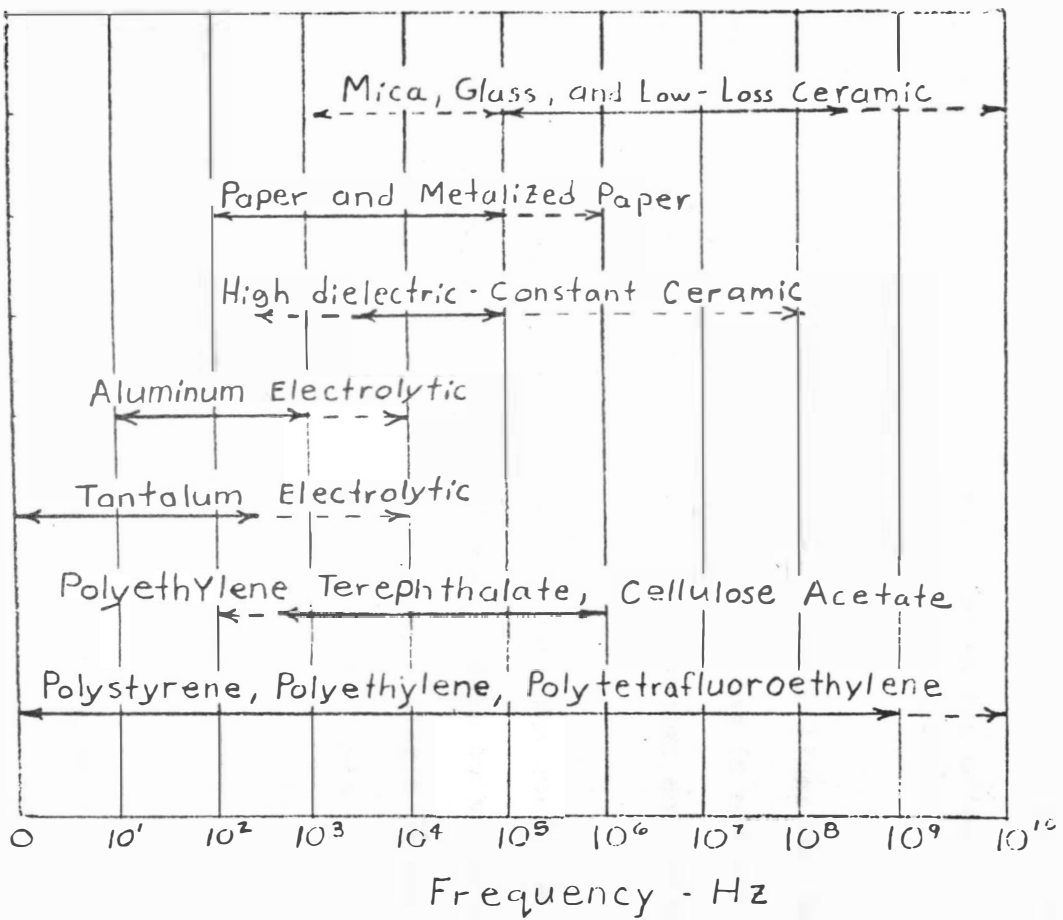
$$L_{int} = \frac{1}{2\pi a \delta \sigma \omega} \text{ h/m.} \quad (2-15)$$

Most component values are determined at a low frequency, about 1,000 Hz, and are not valid at higher frequencies. It may be surprising to know that all capacitors can not be used at high frequencies. At high frequencies dielectric polarization in some capacitors does not have sufficient time to occur; in others the power loss in the dielectric becomes significant. The best dielectrics to use at high frequencies are glass, mica, low-loss ceramic, polystyrene, polyethylene, polytetrafluorethylene (Teflon), and porcelain as shown in Figure 2-10 (10).

It is advisable to use carbon composition resistors with short leads at microwave frequencies. A wire wound resistor will exhibit excessive inductive and capacitive effects.

Inductors should be made from a stiff silver plated wire wound with an air core. A stiff wire is needed to hold the coils rigidly to prevent any change in inductance. Because of skin effect only the area within four skin depths from the surface of the wire needs to be a good current conductor (11). Silver is a good conductor and is commonly used. Air is often used for a dielectric as its dielectric power loss is approximately the power loss of a vacuum.

Because electric and magnetic interaction between adjacent components in



Dashed lines indicate that the capacitor range may be extended by design and manufacturing techniques.

FIGURE 2-10. CAPACITOR DIELECTRIC VERSUS FREQUENCY.

microwave circuits cannot be avoided, it is desirable to keep all parasitic effects constant. The best way of doing this is by mounting the components on a printed circuit board (PC board). This will hold all components and connecting lines rigid, thereby, making all parasitic effects constant.

2-7. Microstrip Design.

Two types of connecting strips or transmission lines may be used on PC boards. The first is stripline and the second is microstrip. In spite of a few adverse properties microstrip is easier and cheaper to construct and is used in this thesis.

Microstrip is a type of planar transmission line consisting of a single strip conductor separated from a ground plane by a dielectric material. It is similar to stripline which has a single conductor sandwiched between two ground planes, with a dielectric material separating the conductor from each ground plane. Microstrip is subject to more fringing and radiation fields than stripline, but proper shielding can control radiation.

Microstrip propagation is not true TEM (Transverse Electromagnetic) mode, because the properties of the dielectric material between the conducting strip and ground plane differ from the properties of the air above the strip. Figure 2-11 is a drawing of a microstrip PC board.

The characteristic impedance of an unshielded microstrip with a zero thickness conductor is (12)

$$Z_0 = \frac{377}{\sqrt{\epsilon_r}} \frac{h}{W} \frac{1}{\left[1 + 1.735(\epsilon_r)^{-0.0724 \frac{W}{h}} - 0.836 \right]} \quad (2-16)$$

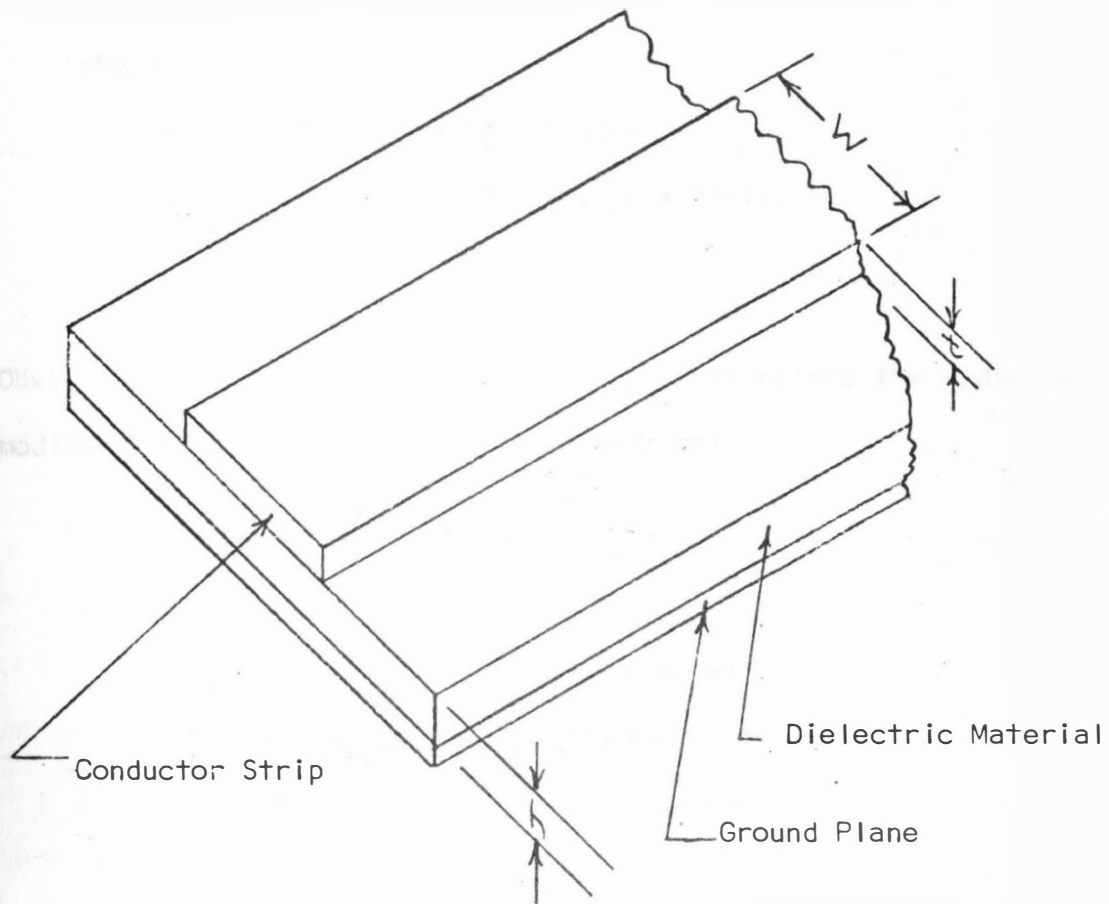


FIGURE 2-11. MICROSTRIP TRANSMISSION LINE.

where

W = width of conductor strip

h = thickness of dielectric material

ϵ_r = dielectric constant.

Obviously the strip has finite thickness, therefore the width must be modified (12).

$$W_{\text{eff}} = W + \frac{t}{\pi} \left[\ln \left(\frac{2h}{t} \right) + 1 \right] \quad (2-17)$$

where

W_{eff} = effective width of conductor

t = thickness of conductor

= 1.4 mils for 1 ounce copper

= 2.8 mils for 2 ounce copper

Because the mode of propagation is not exactly TEM mode, the wavelength, λ_m , of the conductor strip differs from the TEM wavelength, λ_{TEM} . And λ_{TEM} differs from λ_o , wavelength in free space, because the dielectric between the strip and ground plane differs from the air above the strip,

The relationships are (12)

for $\frac{W_{\text{eff}}}{h} \geq 0.6$

$$\frac{\lambda_m}{\lambda_{\text{TEM}}} = \left[\frac{\epsilon_r}{1 + 0.63(\epsilon_r - 1) \frac{W_{\text{eff}}}{h} 0.1255} \right]^{1/2} \quad (2-18)$$

and for $\frac{W_{\text{eff}}}{h} < 0.6$

$$K = \frac{\lambda_m}{\lambda_{\text{TEM}}} = \left[\frac{\epsilon_r}{1 + 0.63(\epsilon_r - 1) \frac{W_{\text{eff}}}{h} 0.0297} \right]^{1/2} \quad (2-19)$$

where

$$\lambda_{\text{TEM}} = \frac{\lambda_o}{\sqrt{\epsilon_r}} \quad (2-20)$$

Experimental results show that the thickness of the conductor strip and ground plane need not be greater than about four skin depths thick (11). The conductor strip can be any good conductor material. The better the conductor the smaller the power losses in the strip.

Some desirable properties of dielectric materials are an $\epsilon_r < 5.0$, low-loss material, high mechanical and electrical strength, low moisture absorption, a constant ϵ_r over a wide range of frequencies, high electrical resistance, low thermal resistance.

Some materials used for printed circuit boards are melamine, epoxy, polyester, phenolic, ceramic and teflon. Fiber glass is frequently combined with a dielectric material to provide mechanical strength. Ceramic and teflon are low-loss materials frequently used in microstrip PC boards. A thick dielectric microstrip PC board is less lossy than a thick dielectric microstrip PC board.

For microwave use, the dielectric constant, conductor strip thickness and the dielectric thickness must be controlled within close tolerances. This increases manufacturing costs over PC boards designed

for low frequency use, which do not have stringent tolerance controls.

Because of the high cost of microwave PC boards Motorola investigated the possibilities of using NEMA (National Electrical Manufacturers Association) designation, type G-10 epoxy-glass laminates for microstrip applications (13). As a result they found over a sampling of six samples from each of eight manufacturers, that the dielectric constant of the laminates varied from 4.61 with a standard deviation of 0.0560 for one manufacture to 5.64 with a standard deviation of 0.1660 for another manufacturer. The variation between the extremes results in a difference of about 1.6 picofarads which can be easily compensated with trimming capacitors between the conductor strip and ground. As the names of the manufacturers were not provided, a median dielectric constant of 5.25 was chosen for this thesis project.

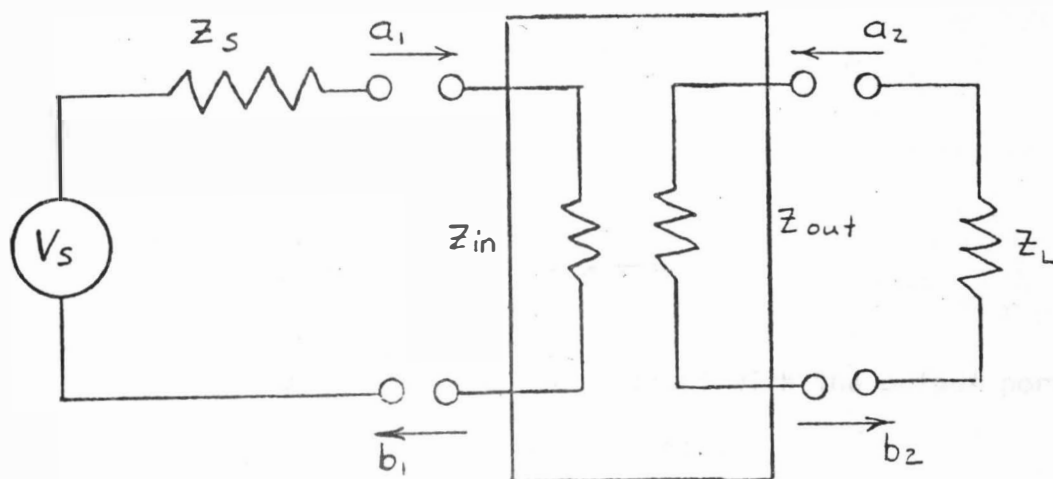
2-8. S-Parameters.

S-parameter circuit design analysis has become the principle tool in microwave design. They are easily and accurately obtained. S, or scattering parameters, are the reflection and transmission coefficients of a 2-port network.

In Figure 2-12, a_1 and a_2 are waveforms incident on the network, while b_1 and b_2 are the waveforms reflected from the network. Coefficients "a" and "b" represent the square roots of power where

$$\text{Power} = \frac{(\text{voltage})^2}{R} . \quad (2-21)$$

If $Z_S = Z_L = R$ than the S_{11} and S_{22} power reflection coefficients become



V_s = Source Voltage

Z_s = Source Impedance

Z_L = Load Impedance

Z_{in} = Input Impedance

Z_{out} = Output Impedance

FIGURE 2-12. S-PARAMETER MODEL.

equal to the voltage reflection coefficients, and S_{12} and S_{21} become a voltage ratio which can be measured with a vector voltmeter.

$$\text{For (14)} \quad b_1 = S_{11}a_1 + S_{12}a_2 \quad (2-22)$$

$$\text{and} \quad b_2 = S_{21}a_1 + S_{22}a_2 \quad (2-23)$$

$$\text{then} \quad S_{11} = \left. \frac{b_1}{a_1} \right| a_2 = 0 \quad (2-24)$$

= Input reflection coefficient with the output port terminated by a matched load.

$$S_{22} = \left. \frac{b_2}{a_2} \right| a_1 = 0 \quad (2-25)$$

= Output reflection coefficient with the input terminated by a matched load.

$$S_{21} = \left. \frac{b_2}{a_1} \right| a_2 = 0 \quad (2-26)$$

= Forward transmission (insertion) gain with the output port terminated in a matched load.

$$\text{and} \quad S_{12} = \left. \frac{b_1}{a_2} \right| a_1 = 0 \quad (2-27)$$

= Reverse transmission (insertion) gain with the input port terminated in a matched load.

The s-parameters can be easily obtained with the use of an 8405A Vector Voltmeter manufactured by Hewlett Packard. Figure 2-13 shows one method of obtaining S-parameters. S_{11} and S_{22} can then be plotted on a Smith Chart to obtain the input and out impedance of the system. Hewlett Packard has several references outlining methods to use in designing microwave amplifiers and oscillators with S-parameters.

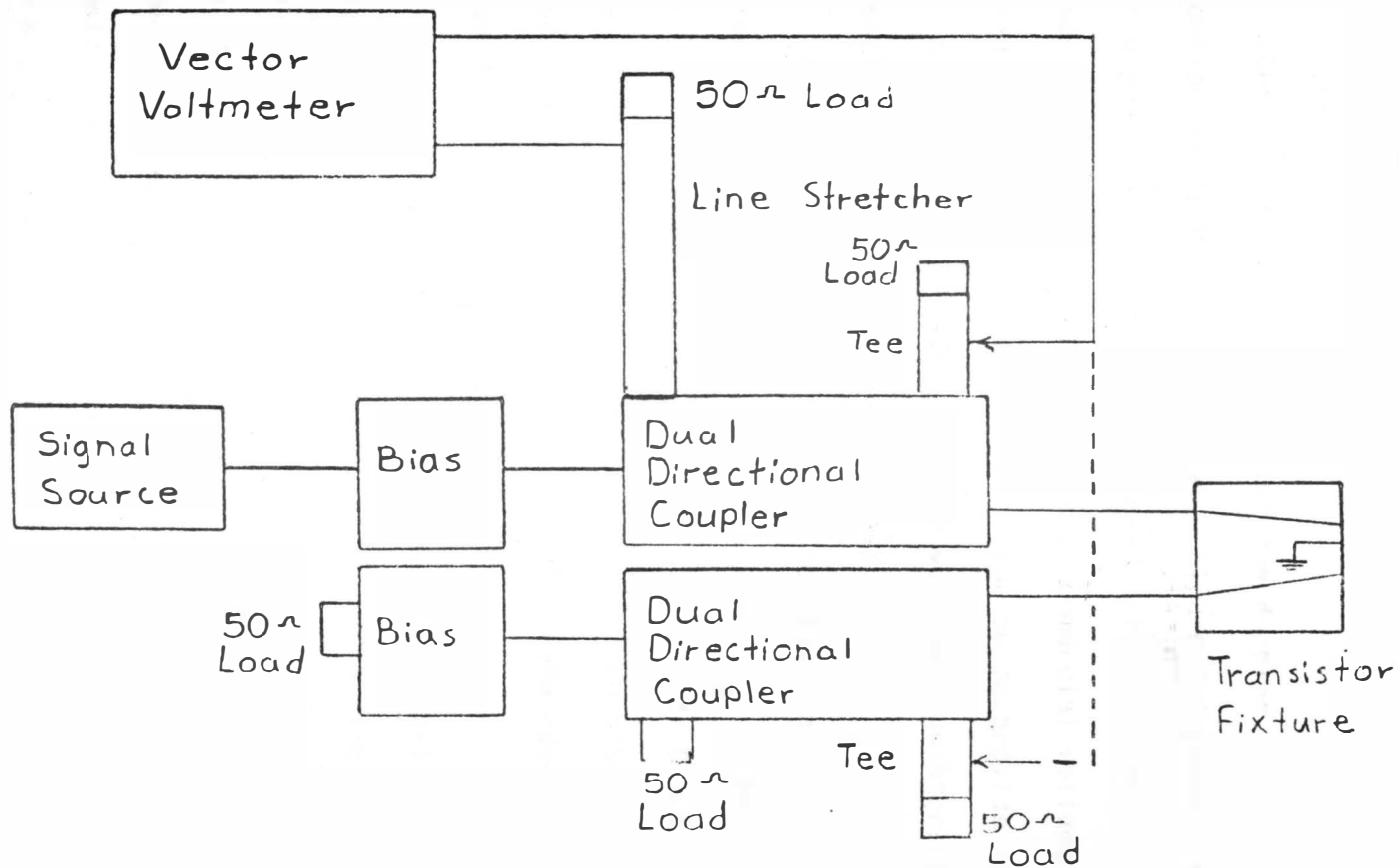


FIGURE 2-13. BLOCK DIAGRAM OF S-PARAMETERS MEASUREMENT SET-UP.

CHAPTER 3

DOPPLER SWITCH.

3-1. Power Supply.

The type of power source for the doppler switch will depend on the location and usage of the system. Figure 3-1 shows a schematic diagram for a power supply that can be used in the doppler switch where 120 Vac power source is available. It uses commercial voltage regulators and operates from a 120 volt a-c power source to provide ± 12 volts and ± 6 volts. Two LM320-5 negative voltage regulators and two LM309 positive voltage regulators are used.

The LM320-5 and the LM309 are connected according to section 7 of reference 15. The variable resistors R_2 , R_4 , R_7 and R_8 are adjusted to provide ± 12 volts and ± 6 volts. The ± 6 volts is used as a power source for the 500 MHz oscillator, the NE565 frequency-to-voltage converter and the TTL logic gates. The remainder of the system operates from the ± 12 volt source.

Gelled lead acid batteries provide an excellent low-noise power source and could be used at locations where line noise is a problem. Because of their portability they are also useful at remote locations, but they will introduce the problem of battery charging. The gelled lead acid battery has the advantage of long storage life, low internal resistance, an ability to be recharged after complete discharge, and it is a completely sealed unit. Disadvantages are cost and the need for

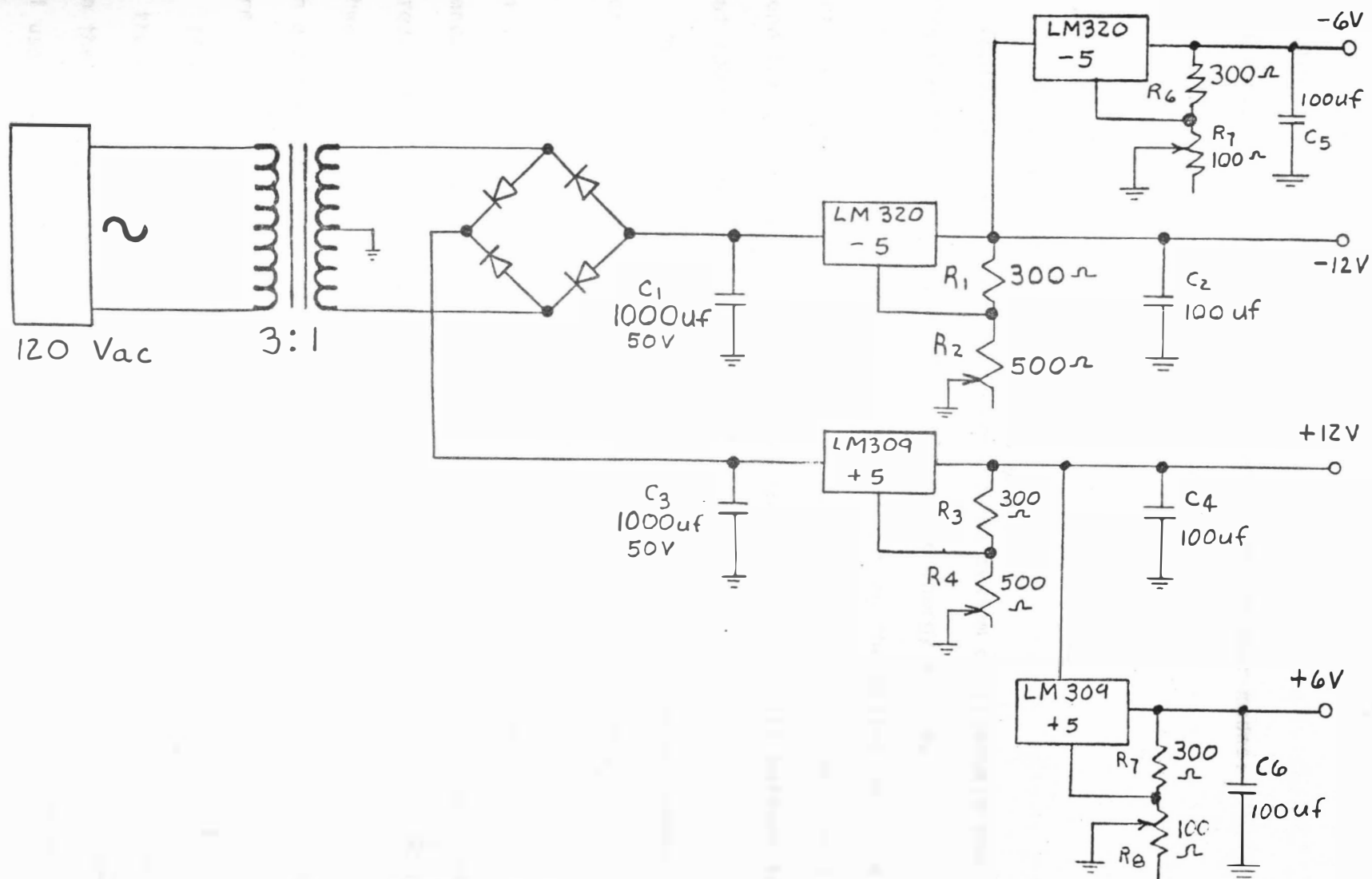


FIGURE 3-1. POWER SUPPLY.

recharging.

The Doppler switch may be operated in one of four modes:

- 1) Continuous
- 2) Pulsed
- 3) Pulsed-on-demand
- 4) Continuous-on-demand.

Continuous mode refers to operating the system continuously and is not recommended for use when a battery is the energy source.

Pulsed mode refers to periodically switching the system on for a short period of time. An on period of .01 second every .1 second will extend the life of a 20 amp-hr, Globe Union, Inc., Gel/Cell battery to about 1000 hours.

Pulsed-on-demand mode refers to turning a system on when needed which will then operate in pulsed mode for the period of demand.

Continuous-on-demand mode is similar to pulsed-on-demand mode except that the transmitter will operate continuously for the period of demand. This mode would work well for locations where low usage is expected. An example would be a location where a private drive enters a highway. A continuous wave transmitter is less expensive to construct than a pulsed transmitter. Therefore, it is more desirable at locations where low usage is expected.

If the indicator is located near an electrical power source it can use the source for power. If there is no electrical source available, then the indicator will have to be battery operated also. The indicator will use more power than the transmitter and the receiver, therefore, it

may be necessary to parallel two or more batteries. The batteries powering the indicator should discharge the same rate as the batteries powering the transmitter and the receiver. Figure 3-2 shows the required battery rating for a desired current drain over a given period of time (16).

A battery powered system has the advantage of being virtually free from 60 Hz line noise. A line operated system has the advantage of being able to supply large amounts of power, except during power outages. Combining these two systems for redundancy, as described below, will provide the best power supply system, but it will be more expensive.

The doppler switch can be powered by four 6 volt batteries connected to a 12 volt d-c source as shown in Figure 3-3. Two batteries in series provide the positive 12 volt source while the other two batteries in series provide the negative 12 volt source with ground being common to both sources. The positive 6 volt source is obtained from the positive side of the first battery above ground potential and the negative 6 volt source is obtained from the negative side of the first battery below ground potential. The manufacturer's data sheet recommends charging Gel-Cell batteries at 2.3 volts per cell therefore a +/- 13.8 volt charging system is used. The charging system should contain a timer programmed to charge the batteries during periods of little use. An example would be a 24 hour charging cycle beginning at midnight and having a five hour charging period. Commercial timers are available in most hardware stores.

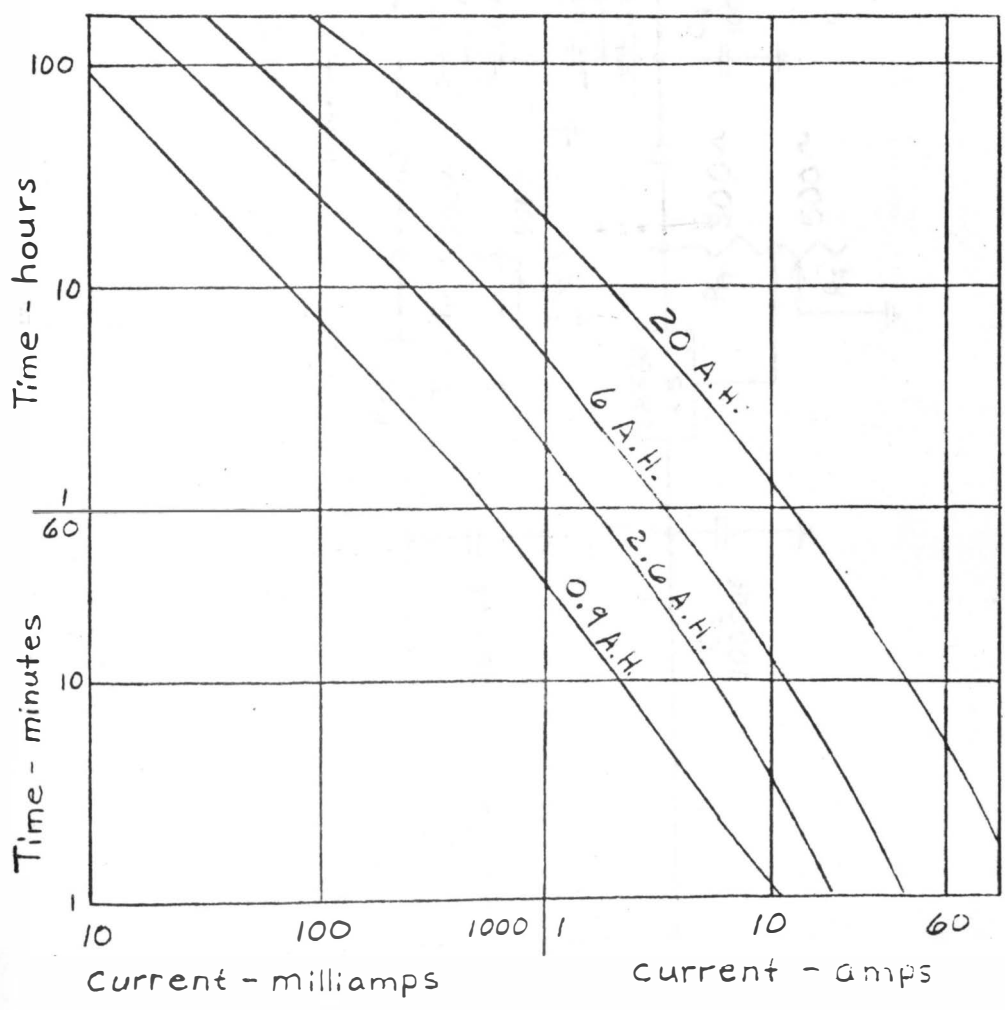


FIGURE 3-2. DESIRED OPERATING TIME VERSUS REQUIRED CURRENT TO DETERMINE BATTERY AMPERE-HOUR CAPACITY.

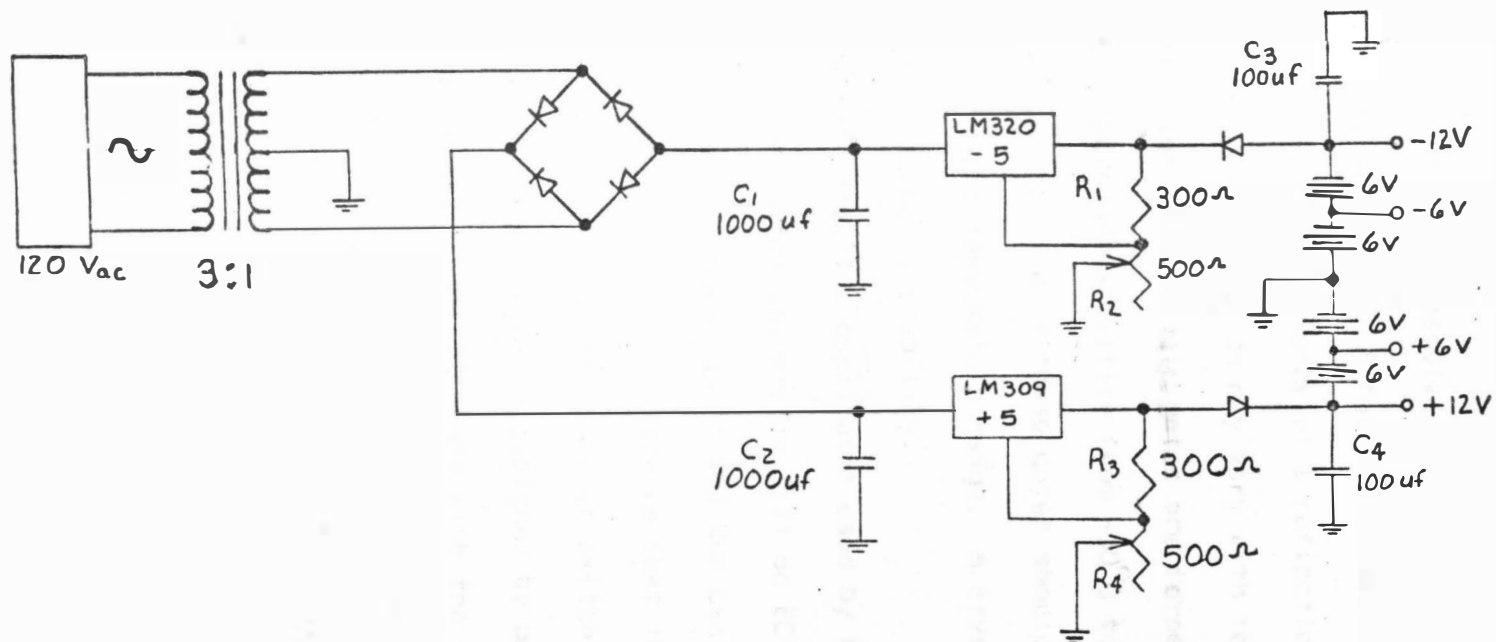


FIGURE 3-3. BATTERY POWER SOURCE WITH CHARGER.

3-2. Oscillator.

The oscillator should be inexpensive, easy to construct and it should have excellent frequency stability. A reflection oscillator meets the first two requirements but the frequency of a reflection oscillator is dependent on its input impedance which may vary with temperature changes. The Motorola MECL III line of ECL logic gates are temperature compensated and have good temperature stability from -30°C to $+85^{\circ}\text{C}$. A reflection oscillator made from one of these gates should have good frequency stability within this temperature range. A crystal may be used to obtain even greater frequency stability.

Figure 3-4 shows a reflection oscillator used by Richard Sawrey in his thesis at South Dakota State University (17) on ECL Gate Oscillators. His data shows that frequencies to 400 MHz can be obtained with MECL II MC1660 OR-NOR Gates. This will require that the second harmonic of a 250 MHz oscillator be used for the doppler switch.

Briefly, a reflection oscillator is designed by obtaining S'_{11} , the input reflection coefficient of the ECL Gate with the amplifiers connected to its output, at the desired frequency. Then by attaching a component or group of components having a reflection coefficient, $K_S = 1/S'_{11}$ to the input of the ECL gates, oscillations of 250 MHz will occur. Because K_S cannot be greater than unity and S'_{11} must be greater than unity, some type of external feedback may be necessary.

Because S'_{11} is a function of the impedance loading the gate the following method should be used to design the oscillator. First, design and build the antenna. Second, design and build the amplifiers. Third,

attach the oscillator of Figure 3-4. Fourth, obtain K_S . Fifth, determine the components to be attached to the input of the gate by plotting K_S on a Smith Chart.

S'_{11} can be calculated by

$$S'_{11} = S_{11} + \frac{S_{12}S_{21}K_L}{1 - S_{22}K_L} \quad (3-1)$$

where

K_L = reflection coefficient of the oscillator loading circuit.

The S-parameters are for the ECL gate and the loading circuit is the amplifiers and the antenna. Direct measurement of S'_{11} , is much easier and should be more accurate as the effect of the interaction of the components will be included in K_S .

3.3 Transmitter.

The design of the transmitter shown in Figure 3-5 is similar to the transmitter designed in reference 12. Class A biasing is used for the design, instead of Class C, though, as the ECL Gate oscillator does not have sufficient power to drive a class C biased transistor. A Motorola type MMT8015 small-signal microwave transistor was selected for the transmitter because it has an f_T of 1.0 GHz, an unneutralized power gain of 15 db ($I_C = 1.0$ mA and $V_{CE} = 6.0$ V). It also has a low noise figure of 2.0 db and a maximum power rating of 200 mW, which is sufficient for a range of 500 ft.

The printed circuit board is restricted to 6 inches by 12 inches,

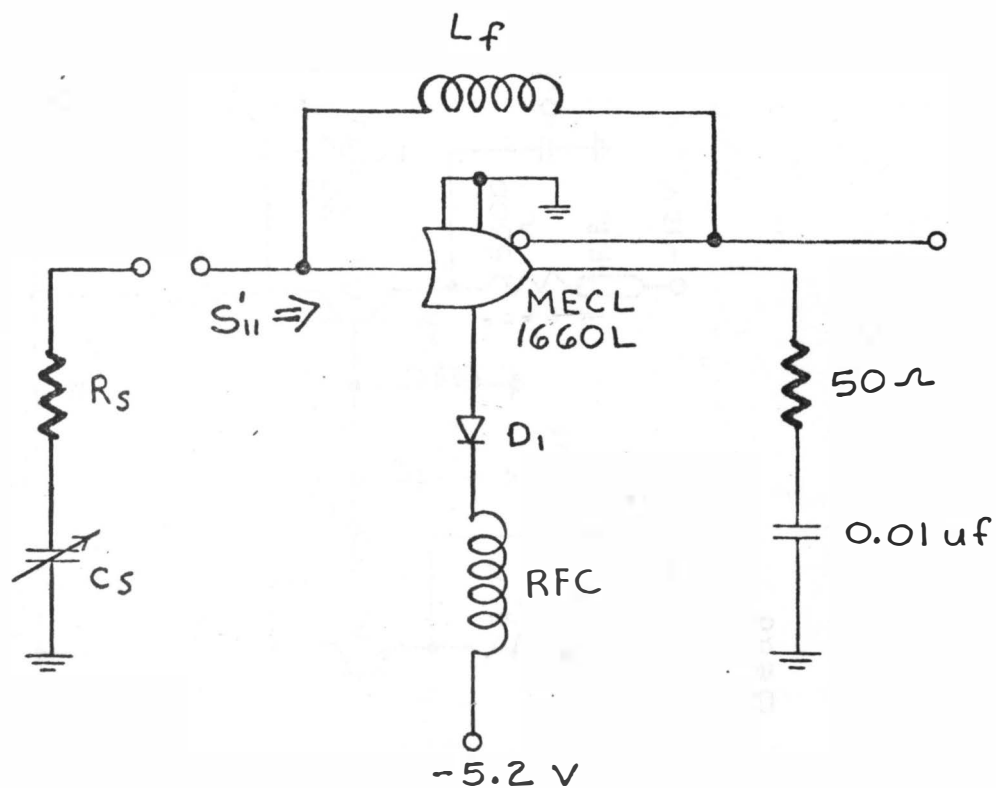


FIGURE 3-4. REFLECTION OSCILLATOR.

because of the dimensions of the mailbox in which it is to be housed.

The transmitter needs sufficient area for an ECL Gate oscillator, three transistors and four sections of microstrip.

It is desirable to have at least three transistors to provide isolation between the oscillator and the antenna otherwise impedance changes at the antenna will affect the oscillator frequency. If the output power is greater than 100 mW after the transmitter is built, C_1 , C_2 , and C_3 can be adjusted to reduce it.

I_{CBO} , the collector to base current with the emitter open, is a function of temperature and can cause the operating point of the transistor to vary with a change in the temperature of the collector-base junction. A measure of a circuit's sensitivity to temperature is the stability factor "S". For common emitter biasing (18)

$$S = \frac{R_E + R_B}{R_E + R_B(1-\alpha)} \quad (3-2)$$

where

R_E = Emitter resistor

and

R_B = Base resistor

The smaller the stability factor the more stable a circuit will be. The ultimate stability factor is '1', which is realized when R_B is equal to zero.

The RF choke in the base circuits of Figure 3-5 has a d-c resistance

near zero and a high impedance at 500 MHz. Therefore, the transmitter amplifiers will have good temperature stability and an input impedance equal to S_{11} .

The d-c bias circuitry for the amplifiers was determined using information from the manufacturer's data sheet. A nominal value of 150 was used for h_{FE} . A power gain of 15 db can be realized with the transistor biased at $V_{CE} = 6.0$ V and $I_C = 1$ mA.

For

$$V_{CC} = +12.0 \text{ V}$$

$$V_{EE} = -6.0 \text{ V}$$

$$I_C = 1.0 \text{ ma}$$

$$h_{FE} = 150$$

$$V_{BE} = 0.7 \text{ V (Silicon)}$$

and

$$V_{CE} = 6.0 \text{ V}$$

the loop equation for the base-emitter circuit is:

$$V_{EE} = V_{BE} + I_E R_E$$

$$6 \text{ V} = 0.7 \text{ V} + R_E (10^{-3} \text{ A}),$$

therefore

$$R_E = 5.3(10^3) \Omega.$$

The loop equation for the collector emitter circuit is

$$V_{CC} + V_{EE} = R_C I_C + V_{CE} + R_E I_C$$

$$12 \text{ V} = 6.0 \text{ V} = R_C (10^{-3} \text{ A}) + 6.0 \text{ V} + 5.3(10^3) \Omega (10^{-3} \text{ A}),$$

therefore

$$R_C = \frac{18.0 - 6.0 - 5.3}{10^{-3}} \frac{V}{A}$$

$$= 6.7 (10^3) \Omega$$

where

$$R_C = \text{collector resistor.}$$

Because the emitter impedance is reflected to the base by a factor of $\beta + 1$, an emitter bypass capacitor is needed. A $0.1 \mu\text{f}$ capacitor will present an impedance of 0.0032Ω in the emitter circuit which will be reflected to the base as 0.48Ω . This is an insignificant value.

The decoupling capacitors and ferrite beads in the supply line are needed to prevent RF frequencies from entering the power supply and feedback which could cause oscillations. The ferrite beads act as an inductance and help suppress current transients.

The S-parameters of the transistor taken from the data sheet, are

$$S_{11} = .7 \angle -94^\circ$$

$$S_{21} = 1.6 \angle +108^\circ$$

$$S_{12} = .065 \angle +40^\circ$$

$$S_{22} = .85 \angle -17^\circ$$

It is desired that each microstrip have a characteristic impedance of 50Ω . Rather than solving for W in equation (2-16) it will be faster

to insert the standard widths of the circuit artwork available for making printed circuit board layouts, and use a calculator to determine which width, if any, will be sufficient. Using equation (2-16) and equation (2-17) with

$$E_r = 5.25 \text{ farad/m}^2$$

$$h = 56.4 \text{ mils}$$

$$W = 125 \text{ mils}$$

and

$$t = 2.8 \text{ mils}$$

$$W_{\text{eff}} = 125 \text{ mils} + \frac{2.8 \text{ mils}}{\pi} \left[\ln \left(\frac{2(56.4 \text{ mils})}{2.8 \text{ mils}} \right) + 1 \right]$$

$$= 129.2 \text{ mils}$$

and

$$Z_o = \frac{377}{\sqrt{5.25}} \times \frac{56.4 \text{ mils}}{129.2 \text{ mils}} \times \frac{1}{\left[1 + 1.735(5.25)^{-0.0724} \left(\frac{129.2 \text{ mils}}{56.4 \text{ mils}} \right)^{-0.836} \right]}$$

$$= 40.6 \Omega$$

This is too low. For $W = 93 \text{ mils}$

$$W_{\text{eff}} = 93 \text{ mils} + \frac{2.8 \text{ mils}}{\pi} \left[\ln \left(\frac{2(56.4 \text{ mils})}{2.8 \text{ mils}} \right) + 1 \right]$$

$$= 97.2 \text{ mils}$$

and

$$Z_o = \frac{377}{\sqrt{5.25}} \times \frac{56.4}{97.2} \times \frac{1}{\left[1 + 1.735(5.25)^{-0.0724} \left(\frac{97.2}{56.4} \right)^{-0.836} \right]}$$

$$= 48.3 \Omega$$

Normalized to 50 Ω

$$Z = \frac{48.3}{50} = .966,$$

which gives a reflection coefficient (4)

$$\begin{aligned} K_R &= \frac{Z - 1}{Z + 1} \\ &= \frac{.966 - 1}{.966 + 1} \\ &= -.017 \end{aligned}$$

k; then, is $.017 \angle 180^\circ$ which is insignificant, so Z_0 will be considered to be 50 Ω .

The wavelength of the signal in the dielectric, using equation (2-18) and equation (2-20) is

$$\begin{aligned} \lambda_{\text{TEM}} &= \frac{3(10^8)\text{m/sec}}{5(10^8)\text{Hz}} \times \frac{1}{\sqrt{5.25}} \\ &= .262 \text{ m} \\ \lambda_{\text{TEM}} &= 26.2 \text{ cm.} \end{aligned}$$

For

$$\begin{aligned} K &= \left[\frac{5.25}{1 + 0.63(5.25-1) \left(\frac{128.3}{56.4} \right)^{.1225}} \right]^{1/2} \\ &= 1.151 \end{aligned}$$

then

$$\begin{aligned}\lambda_m &= \lambda_{\text{TEM}}^K \\ &= 26.2 \text{ cm}(1.151) \\ &= 30.16 \text{ cm.}\end{aligned}$$

Smith Charts were used to determine the proper impedance needed to match the input of A_1 and the output of A_3 to 50Ω . They were also used to match the outputs of A_1 and A_2 to the inputs of A_2 and A_3 , respectively (4, 12).

Figure 3-6, shows the steps in matching S_{11} , the input of A_1 , to 50Ω . The input impedance of A_1 normalized to 50Ω is $0.33 - j0.9 \Omega$. Moving $0.394 \lambda_m$ from A_1 towards the generator to Point B, maintaining a constant SWR, results in an effective impedance of $0.22 + j0.5 \Omega$. The length of the microstrip needed to accomplish this is

$$\begin{aligned}L_{\text{ms}} &= 0.394(30.16 \text{ cm}) \frac{1 \text{ in}}{2.54 \text{ cm}} \\ &= 4.7 \text{ inches.}\end{aligned}$$

Moving 180° around the Smith Chart from Point B to Point C, maintaining a constant radius from 1.0, converts from impedance coordinates to admittance coordinates. A dashed circle representing the inverse of the points on the impedance unity circle is drawn on the Smith Chart. Positive susceptance (parallel capacitance) of 1.05γ is added

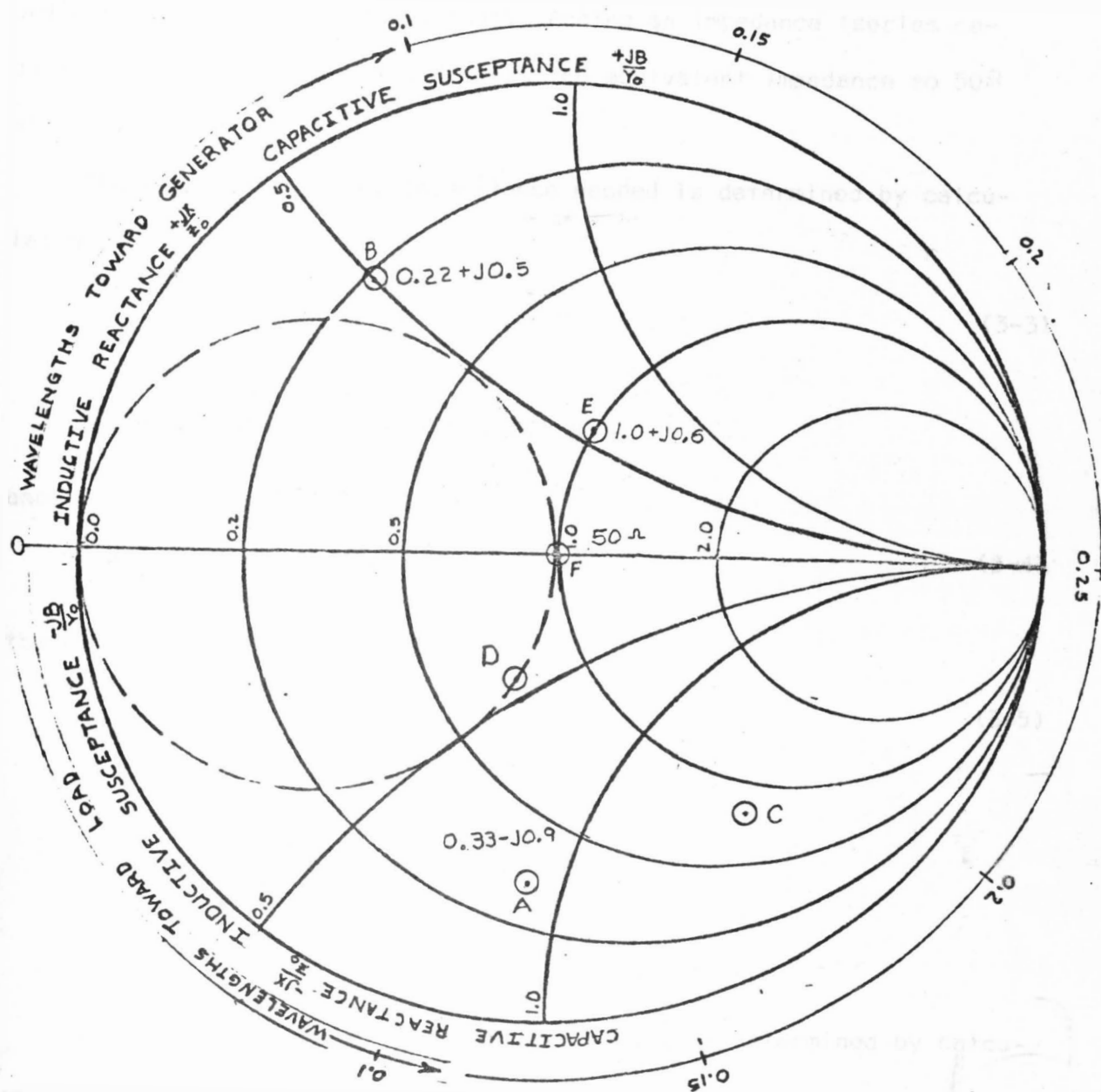


FIGURE 3-6. MATCHING S_{11} to 50 OHMS.

to point C to reach point D on the unity admittance circle. Rotating Point D 180° around the chart to Point E converts the Smith Chart coordinates back to impedance values. Adding an impedance (series capacitance) of $-j0.6$ to Point E moves the equivalent impedance to 50Ω at Point F.

The value of parallel capacitance needed is determined by calculating

$$Y_o = \frac{1}{Z_o} = \frac{1}{50 \Omega} = 0.02 \text{ } \mathcal{V} \quad (3-3)$$

$$B_c = 0.02(1.25 \text{ } \mathcal{V}) = 0.025 \text{ } \mathcal{V}$$

and

$$X_{cp} = \frac{1}{B_c} = 40.0 \Omega. \quad (3-4)$$

therefore

$$C_p = \frac{1}{2\pi f X_c} \quad (3-5)$$

$$= \frac{1}{2\pi(5)10^8(40)}$$

$$= 8.0 \text{ Pf.}$$

The value of the series capacitance needed is determined by calculating

$$X_{cs} = 0.6(50) = 30 \Omega$$

$$C_s = \frac{1}{2\pi(5)10^8(30)}$$

$$= 10.6 \text{ Pf.}$$

Similar steps were made to match S_{22} to S_{11} and S_{22} to 50Ω , respectively.

The results for matching S_{22} to S_{11} are

$$L_{ms} = 7.1 \text{ in}$$

$$C_p = 9.4 \text{ pf}$$

$$C_s = 5.0 \text{ pf}$$

and the results for matching S_{22} to 50Ω are

$$L_{ms} = 7.1 \text{ in}$$

$$C_p = 6.4 \text{ pf.}$$

Only a parallel capacitor was needed to match S_{22} to 50Ω but a series capacitor is needed to block d-c. A $0.01 \mu\text{f}$ capacitor is used for d-c blocking as it will present an impedance of only

$$\begin{aligned} X_c &= \frac{1}{2\pi(5)10^8(10^{-8})} \\ &= 0.03 \Omega \end{aligned}$$

which will not affect the circuit to any degree.

Figure 3-7 shows how the transmitter would be laid out on the printed circuit board. The board is G-10 epoxy glass laminate with one-ounce copper on each side. The side opposite the microstrip conductors must be a solid copper ground plane.

3-4. Antenna.

Figure 3-8 shows a halfwave dipole antenna inside of a standard

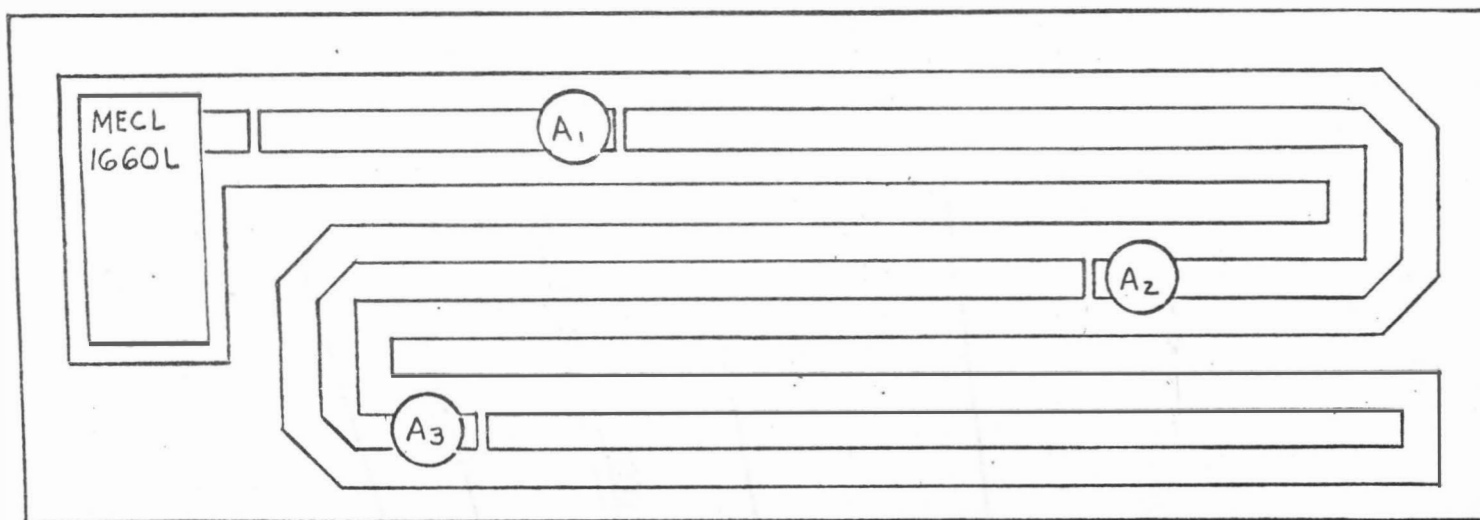


FIGURE 3-7. PRINTED CIRCUIT BOARD LAYOUT FOR TRANSMITTER.

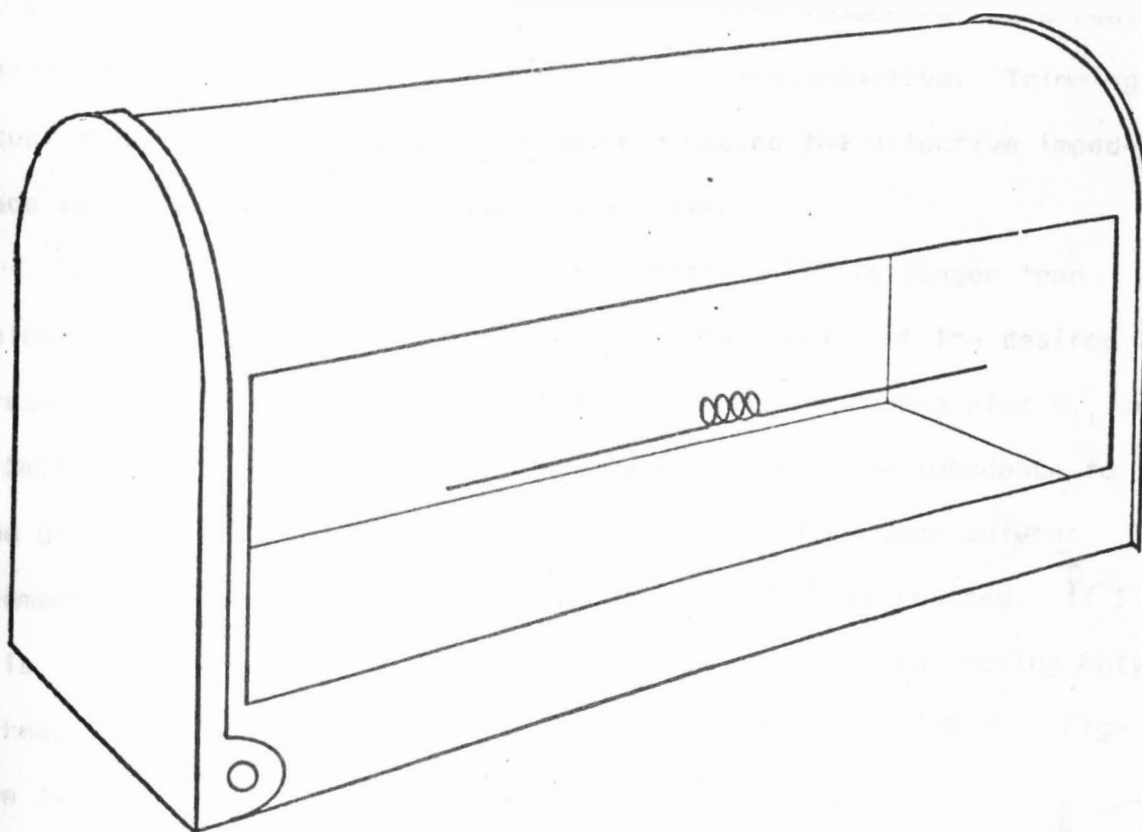


FIGURE 3-8. MAILBOX ENCLOSURE WITH ANTENNA.

rural mailbox. A rectangular hole 5.5 inches by 17 inches is cut in the side of the mailbox to permit the passage of electromagnetic radiation. The antenna is then mounted in the center of the rectangular hole with the back of the mailbox used as a reflector. An RF choke is used between the elements of the antenna to short any low frequencies to ground while presenting a high impedance to 500 MHz.

Because the sides of the enclosure affect the impedance of the antenna a method was needed for impedance matching the antenna to the 50 Ω transmission line. In this case the antenna impedance had a real part close to 50 Ω and an imaginary part that was inductive. Trimming equal amounts from each end of the antenna caused the effective impedance to be less inductive, or more capacitive.

The design technique is to cut an antenna a little longer than calculated results indicate, obtain S_{11} of the antenna at the desired frequency with the antenna inside the mailbox enclosure and plot S_{11} on a Smith chart. Then use discrete components to move the impedance to the unity circle. Then successively cut 1/8 inch from each antenna element and record S_{11} after each cutting until 50 Ω is reached. If 50 Ω is not reached once the real axis is crossed, a component having only a real value may be added in series or parallel to obtain 50 Ω . Figure 3-9 shows the results of each step.

For optimum radiation of energy the length of a dipole antenna should be one-half wavelength. The wavelength for a frequency of 500 MHz is

$$\begin{aligned}\lambda &= \frac{c}{f} = \frac{3(10^8)\text{m/sec}}{5(10^8)\text{HZ}} & (3-6) \\ &= 0.6 \text{ meter}\end{aligned}$$

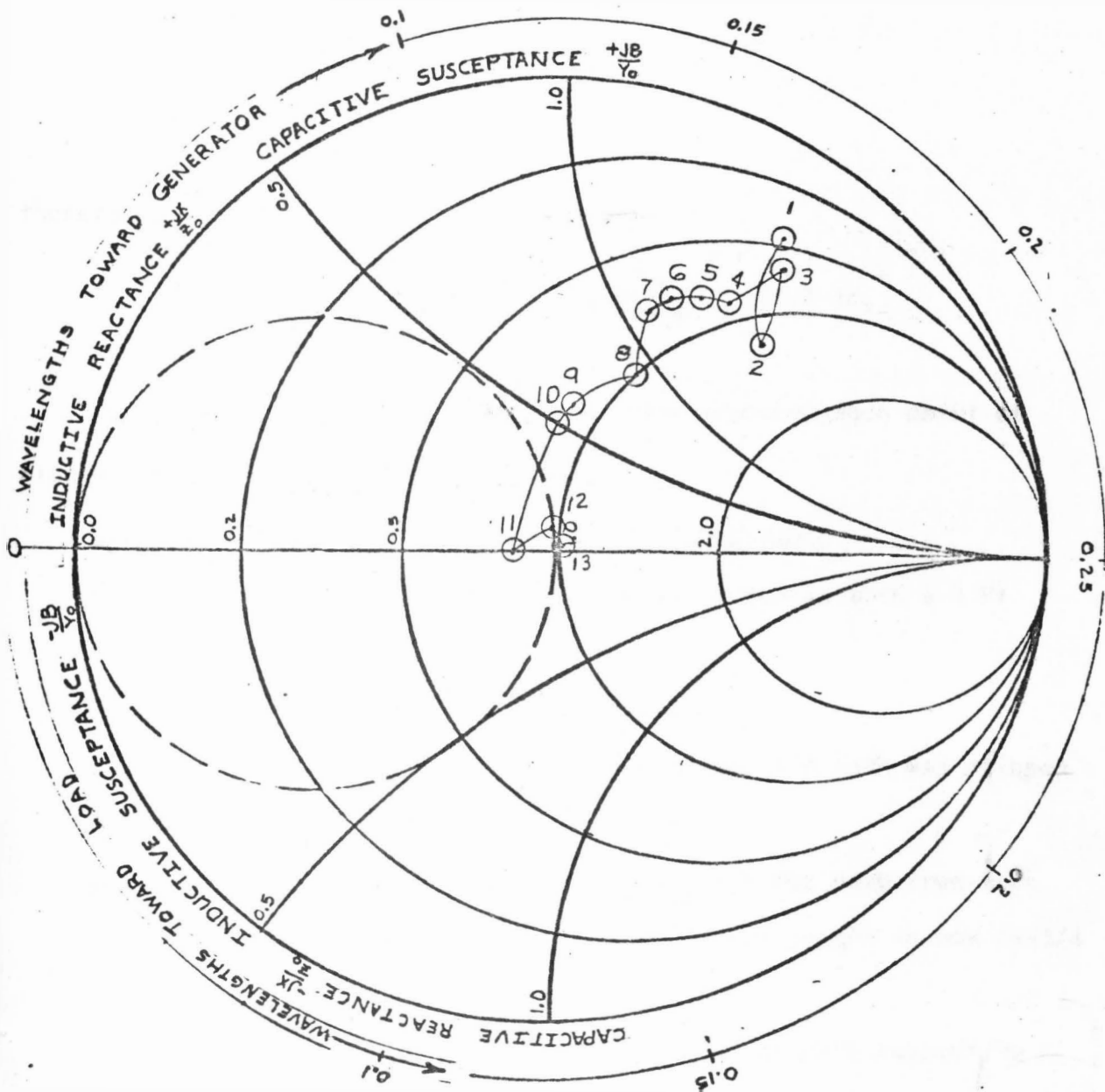


FIGURE 3-9. TRIMMING ANTENNA TO 50 Ω .

where

λ = wavelength

C = Propagation velocity

= speed of light

f = frequency

therefore,

$$\frac{\lambda}{2} = \frac{0.6 \text{ meter}}{2} \cdot \frac{100 \text{ cm}}{1 \text{ m}} \cdot \frac{1 \text{ in}}{2.54 \text{ cm}} = 11.8 \text{ in.}$$

The points below explain the process that produced each point of Figure 3-9.

- Point #1 The antenna was cut to 13-3/4 inch,
- Point #2 A 20 Ω resistor was added in series with a 5 Pf capacitor.
- Point #3 The capacitor was removed.
- Point #4 The capacitor was replaced and 1/8 inch was clipped from each end of the dipole.
- Points #5 - #11 Successive removal of 1/8 inch was made from each end of the dipole. The antenna length is now 11-3/4 inch.
- Point #12 The components were repositioned with respect to each other.
- Point #13 Another 5 Pf capacitor was added in series to make a total of 2.5 Pf.

The antenna impedance is now approximately 50Ω and it will match the impedance of the RG 58 A/U antenna cable.

Data was taken with the antenna installed in the mailbox enclosure, which simulates a flat reflector, to determine its beamwidth. The results, Table 3-1, show a good symmetrical beam having a 60° beamwidth. The width of the beam at 500 feet is 600 feet, which is considered adequate for the application. Too wide a beam would disperse the energy too rapidly and too narrow a beam would make installation alignment critical, therefore, difficult. A Hewlett Packard 3406A Sampling Voltmeter was used for the measurements.

3.5 Receiver, Amplifiers and Filter.

The receiver is shown in Figure 3-10 and 3-11. It is designed to amplify the difference between the transmitted signal and the doppler frequency shifted signal. Referring to Equation 2-8,

$$S_{\min} = \frac{P_t G^2 \lambda^2 \sigma}{(4\pi)^3 R_{\max}^4}, \quad (3-7)$$

P_t is 0.1 watt and λ is 0.6 meter or 1.97 feet. The maximum range has been determined to be 500 feet. The effective antenna area, A_e , as determined in chapter two, is 0.5 square feet. The area of the window in the mailbox is 0.65 square feet. The effective area of the antenna is less than the area of the window, therefore, the window will have negligible effect on the effective antenna area. From Section 2-3 a dipole antenna with a flat reflector one-quarter wavelength from the antenna will have a power gain of four (6 db). This condition is

TABLE 3-1. ANTENNA BEAMWIDTH (500 MHz) - NORMALIZED VALUES.

<u>Degrees from Boresight (0°)</u>	<u>Voltage Normalized to value at 0°</u>	
	<u>Left of Boresight</u>	<u>Right of Boresight</u>
0	1	1
5	0.97	0.97
10	0.94	0.94
15	0.91	0.91
20	0.85	0.85
25	0.82	0.82
30	0.74	0.74
35	0.68	0.68
40	0.62	0.65
45	0.53	0.56
50	0.41	0.50
55	0.38	0.41
60	0.32	0.38
65	0.26	0.32
70	0.24	0.26
75	0.21	0.24
80	0.18	0.18
85	0.16	0.19
90	0.16	0.18

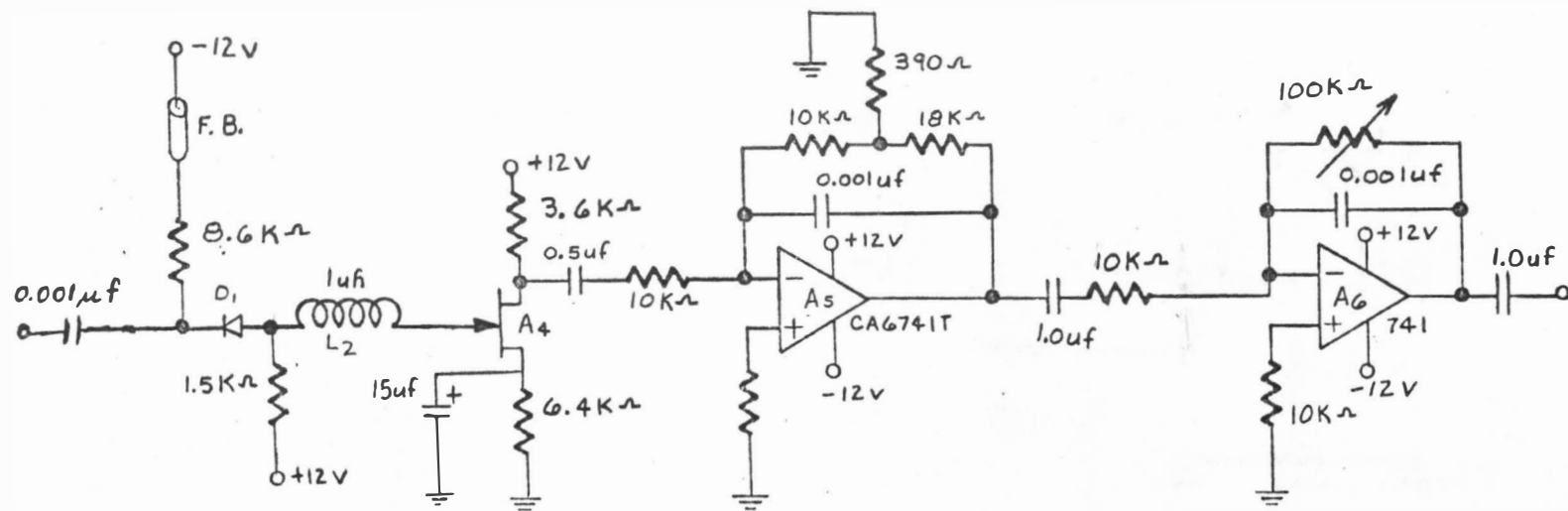


FIGURE 3-10. RECEIVER.



approximated with the dipole in the mailbox enclosure.

The effective area of an automobile for 500 MHz was not given in any of the references checked. An approximation for σ was obtained from reference 2 as follows: For a flat reflector having an area 'A'

$$\sigma = \frac{4\pi A^2}{\lambda^2}$$

For a flat reflector having the dimensions of a midsize automobile, approximately four feet by five feet,

$$\sigma = \frac{4\pi (20 \text{ ft.}^2)^2}{(1.97 \text{ ft})^2}$$

$$= 1,295 \text{ ft.}^2$$

For a small airplane σ was given as 200 ft.^2 . A midsize automobile would be expected to fall between these two values. To be conservative 200 ft.^2 was used. The actual value of σ could be empirically determined for 500 MHz.

Substituting the above values into equation (3-7) gives

$$S_{\min} = \frac{.1 \text{ watt}(4)^2(1.97 \text{ ft})^2(200 \text{ ft}^2)}{(4\pi)^3 (500 \text{ ft})^4} = \frac{1,242 \text{ ft}^4 - \text{watt}}{(1.24)(10^{14}) \text{ ft}^4}$$

$$= 10^{-11} \text{ watt.}$$

For an antenna impedance of 50Ω ,

$$P_R = E_R^2 / Z_A \quad (3-8)$$

and

$$E_R = \sqrt{P_R Z_A},$$

then

$$\begin{aligned} E_R &= \sqrt{10^{-11} \text{ W}(50\Omega)} \\ &= 22.4(10^{-6}) \text{ volts} \end{aligned}$$

where

P_R = received power

Z_A = antenna impedance

E_R = voltage of received signal.

From reference (19) adding two signals, V_c and V_m , will not result in a signal having a frequency equal to the difference between the original two signals until this sum has passed through a nonlinear device such as a diode.

A diode is a nonlinear device having a square-law characteristic. If two signals are added together and passed through a diode the sum will be squared. It can be shown mathematically that a signal having a

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A diode is a nonlinear device having a square-law characteristic. If two signals are added together and passed through a diode the sum will be squared. It can be shown mathematically that a signal having a

frequency equal to the difference between the original two frequencies plus several other frequencies, will result. A filter can then be used to select the signal having the desired frequency.

Adding the signals

$$V_c = V_{cm} \cos \omega_c t \quad (3-9)$$

and

$$V_m = V_{mm} \cos \omega_m t \quad (3-10)$$

results in

$$V = V_c + V_m.$$

Passing V through a diode square-law detector gives

$$V^2 = (V_c + V_m)^2$$

or

$$V^2 = (V_{cm} \cos \omega_c t)^2 + 2V_{cm} V_{mm} \cos \omega_c t \cdot \cos \omega_m t + (V_{mm} \cos \omega_m t)^2.$$

A trigonometric expansion results in

$$\begin{aligned} V^2 = & \frac{V_{cm}^2}{2} + \frac{V_{cm}^2}{2} \cos 2\omega_c t + V_{cm} V_{mm} \cos(\omega_c + \omega_m)t + \\ & V_{cm} V_{mm} \cos(\omega_c - \omega_m)t + \frac{V_{mm}^2}{2} + \frac{V_{mm}^2}{2} \cos 2\omega_m t. \end{aligned} \quad (3-11)$$

Proper filtering will result in

$$V' = V_{cm} V_{mm} \cos(\omega_c - \omega_m)t. \quad (3-12)$$

Which is the frequency shift due to the doppler phenomenon. The magnitude is

$$|V^I| = V_{cm} V_{mm}.$$

It was determined experimentally that the phased locked loop used as a frequency-to-voltage converter will not lock on to a signal less than 10 millivolts in amplitude. Therefore, the receiver must provide a minimum voltage gain of

$$A_v(\min) = \frac{10\text{mv}}{V_{cm} V_{mm}}$$

For

$$\begin{aligned} V_t &= V_{cm} \\ &= \sqrt{P_t Z_o} \\ &= \sqrt{.1\text{w} (50\Omega)} \\ &= \sqrt{5} \\ &= 2.2 \text{ volts} \end{aligned}$$

where

$$V_t = \text{voltage transmitted}$$

$$P_t = \text{Power transmitted}$$

$$Z_o = \text{Characteristic impedance of line.}$$

For

$$E_R = V_{mm}$$

$$\begin{aligned}
 A_v(\min) &= \frac{10^{-2}}{2.2(22.4) \cdot 10^{-6}} = \frac{10^{-2}}{49.3(10^{-6})} \\
 &= 203
 \end{aligned}$$

and

$$V_{cm} V_{mm} = 49.3(10^{-6}) \text{ volts.}$$

In the transmitter of Figure 3-5 all low frequencies located on the antenna side of C_4 will be shorted to ground by L_1 . The transmitted and received signals are blocked by L_2 in the receiver, Figure 3-10. The receiver is connected to the transmitter cable on the transmitter side of C_4 of Figure 3-5 through D_1 and L_2 , as the antenna transmission line is used to sum the transmitted and received signals. Diode D_1 performs the squaring operation F_D , the difference between the transmitted and the received signals, will be amplified by A_4 , A_5 and A_6 .

A field-effect transistor (FET) was used as the first stage in the receiver because FETs typically have lower noise figures than operational amplifiers (op amps). The SN5460 FET used as A_4 has a noise figure of 2.0. This will provide an initial signal-to-noise ratio which is better than could be obtained with an op amp.

A_4 provides a voltage gain of 8, and A_5 provides a voltage gain of 42. A_5 is an RCA 6741T low-noise op amp which has a noise level of about 4 μ V referred to the input. A_6 is a commercial grade 741 op amp. It has a variable resistor in its feedback loop which can vary the gain of A_5 from near zero to 10. A maximum receiver voltage gain of 3,360

is possible. The reason for a gain control is to permit a user to select a maximum range most suitable for the application. Op amps A_7 and A_8 , of figure 3-11 are connected as a bandpass filter to pass frequencies from 45 Hz to 120 Hz which correspond to 30 mph and 80 mph respectively. R_2 adjusts the center frequency of the filter while R_1 adjusts its bandwidth. Table 3-2 shows the results of testing the filter and frequency-to-voltage converter outputs versus frequency.

3-6. Frequency-To-Voltage Converter.

A phased locked loop (PLL) is used as a frequency to voltage converter. The advantage of a PLL is that it will lock onto a signal with a small amplitude. Another advantage is that although the minimum detectable signal input is 10 mv rms, the PLL will lock onto it even if the noise level is of the same amplitude. A_9 has a variable feedback resistor, R_4 , which allows the PLL output voltage to be amplified to a level useable by the sample and hold network. R_4 can be adjusted to provide voltage gains from near zero to 30.

Letting the output of A_9 swing negative reduces the bias voltage on Q_{10} and Q_{11} of the sample and hold network during the hold mode, causing them to begin conducting. This is avoided when there is no signal input to the receiver by keeping the output of A_9 positive during this time. To calibrate R_3 and R_4 the input to the receiver is grounded and R_3 is adjusted until the output of A_9 equals zero volts. Then, with a signal equivalent in amplitude to the received signal expected for the desired range (22.4 μ V for 500 feet). R_4 is adjusted so that the output

of A_g varies over the range from -10 volts to +10 volts as the frequency of the input signal is changed from 45 Hz to 94 Hz.

From equation (2-4) it can be seen that a vehicle approaching the radar will generate an f_{doppler} larger than f_{TX} , while a vehicle moving away from the radar will generate an f_{doppler} less than f_{TX} . Therefore, approaching vehicles will cause F_D to decrease in frequency with respect to time (resulting in an increasing PLL output voltage), while receding vehicles will cause F_D to increase with time (resulting in a decreasing PLL output voltage). The reason for this is that the vehicle's path is not coincident with a line from the vehicle to the radar set; therefore, the radial velocity, and also the doppler frequency, will be a function of the actual velocity and the distance between the vehicle and the radar set as shown in Figure 3-12.

The desired output of A_g , therefore, will be to decrease with increasing frequency and increase for decreasing frequencies as shown in Figure 3-13. The output of A_g is then processed by the direction detector. The actual output voltages for A_g at various input frequencies are listed in Table 3-2 and are plotted in Figure 3-14.

3-7. Direction Detector.

It is desirable to have the system lock onto just those vehicles approaching the radar set not moving away from it. Figure 3-15 is a schematic of the sample and hold network which detects the vehicle's direction by determining whether the doppler frequency is increasing or decreasing in value (20). It does this by comparing its previous

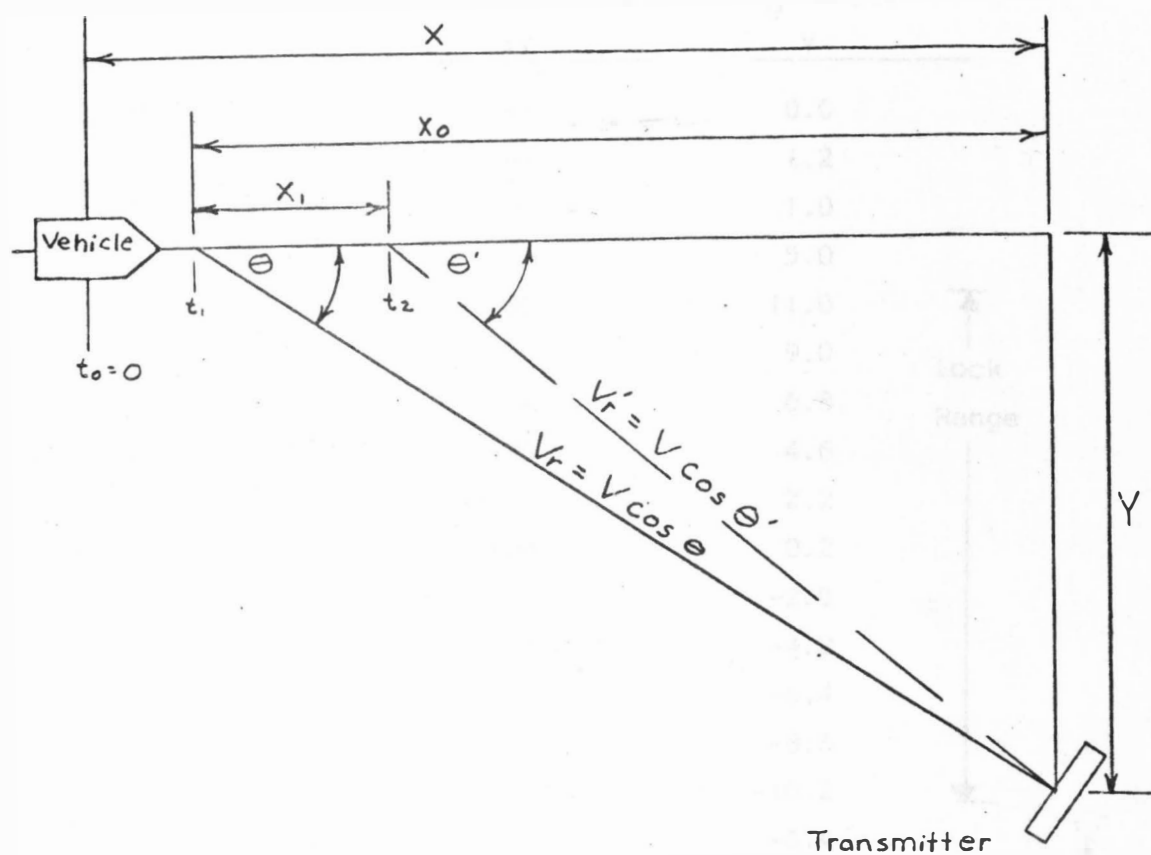


FIGURE 3-12. RADIAL VELOCITY.

TABLE 3-2. FREQUENCY INTO A_1 AND CORRESPONDING VOLTAGE OUT OF FILTER
AND PHASED LOCKED LOOP WITH V_{in} EQUAL TO $30 \mu V$.

Frequency Hz	Filter Output mV	A_9 Output V
25	--	0.0
30	--	1.2
35	--	1.0
40	40	5.0
45	50	11.0
50	55	9.0
55	--	6.8
60	70	4.6
65	--	2.2
70	120	0.2
75	--	-2.0
80	190	-4.2
85	--	-6.4
90	180	-8.6
94	--	-10.2
95	--	-6.4
100	130	-2.2
110	90	0.0
120	70	
130	60	
150	30	



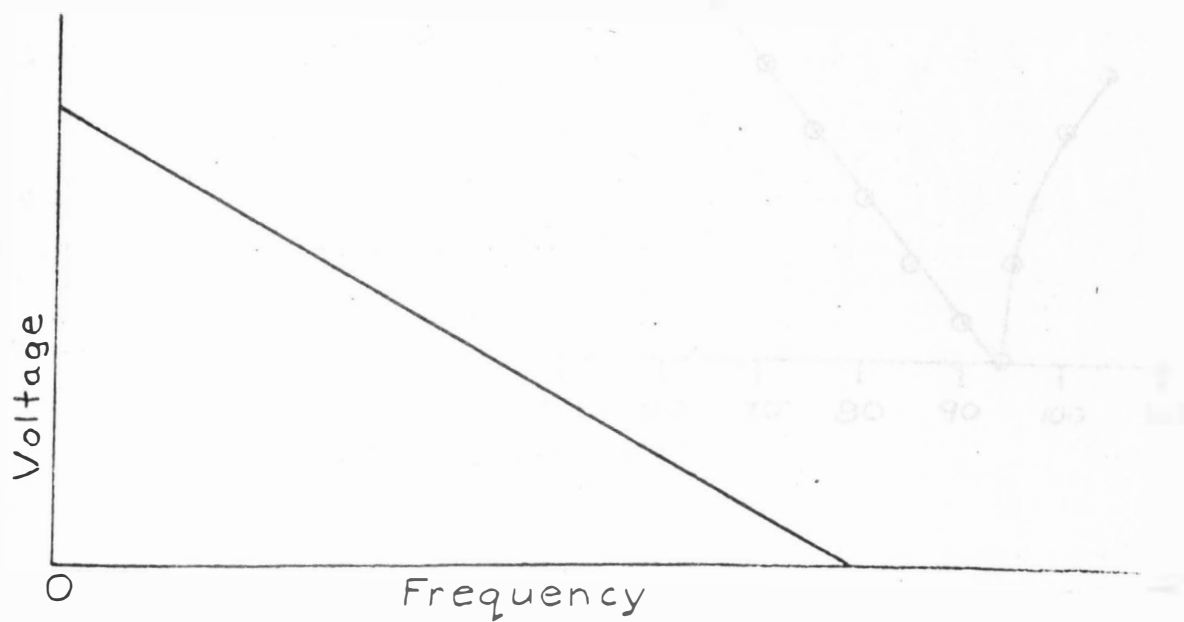


FIGURE 3-13. DESIRED FREQUENCY-TO-VOLTAGE RESPONSE.

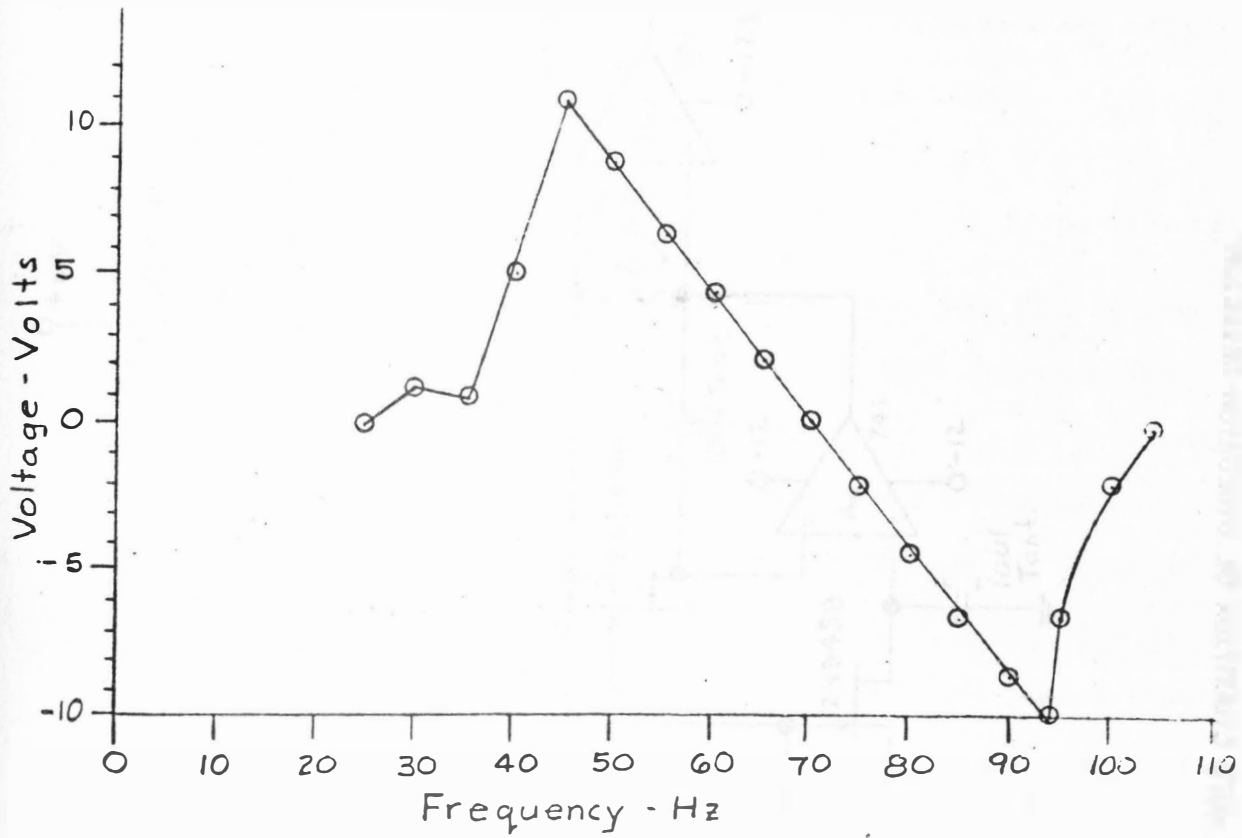


FIGURE 3-14. ACTUAL FREQUENCY-TO-VOLTAGE RESPONSE.

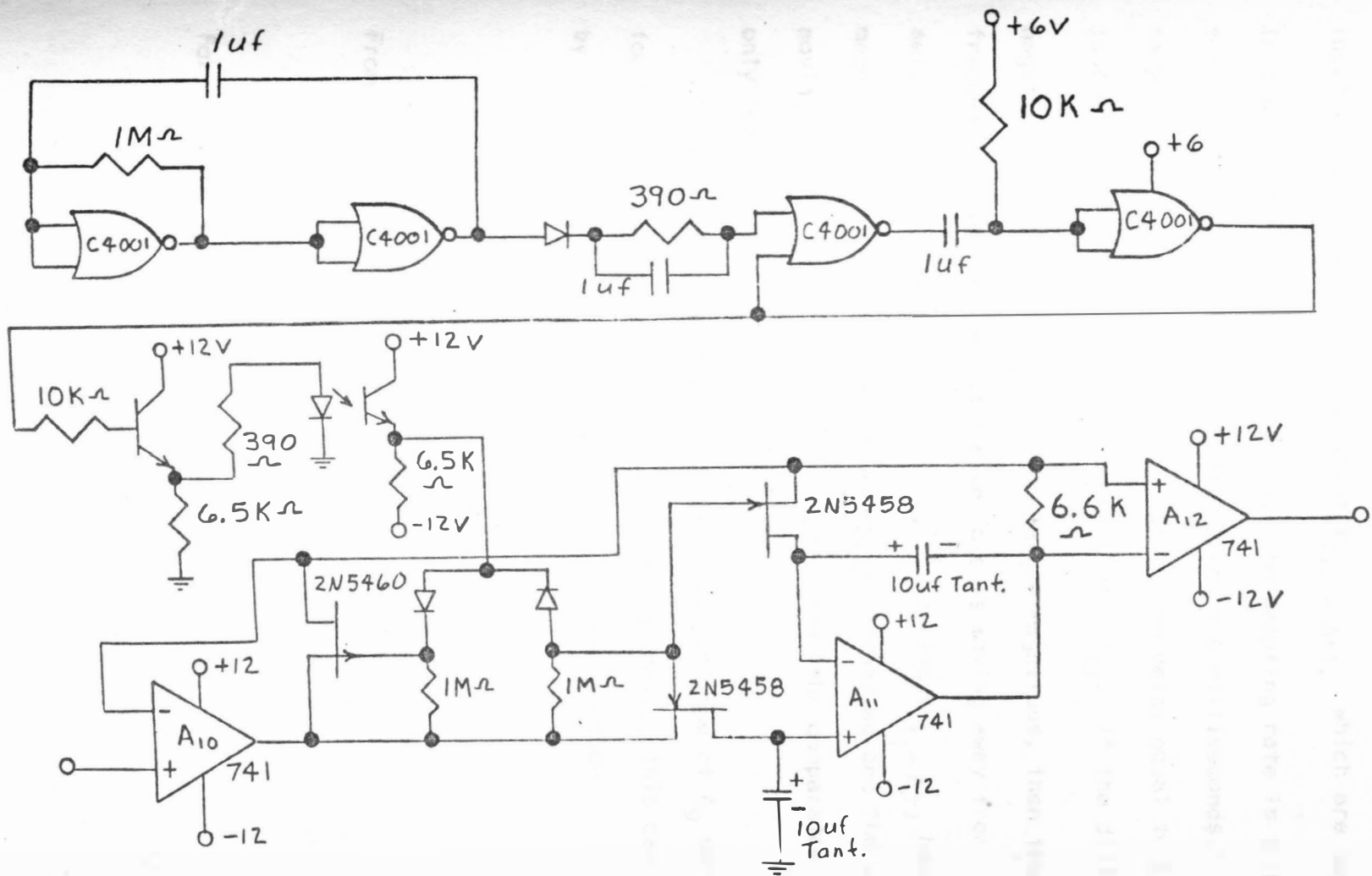


FIGURE 3-15. SAMPLE AND HOLD CIRCUITRY OF DIRECTION DETECTOR.

input $V_{(t_1)}$ with its present input $V_{(t_1 - \Delta t)}$, which are separated in time by Δt , the sampling period. The sampling rate is 1 Hz, therefore, Δt is 1 second. The sampling time is 6 milliseconds. $V_{in}(t)$ is sampled and held as $V_{(t_1)}$, then after a time delay equal to Δt , $V_{(t_1)}$ is subtracted from $V_{(t_1 - \Delta t)}$ in comparator A_{12} . If the difference is negative, indicating $V_{(t_1)}$ has the larger magnitude, then the doppler frequency is increasing and the vehicle is moving away from the radar set. If the difference is positive, indicating $V_{(t_1 - \Delta t)}$ has the larger magnitude, then the doppler frequency is decreasing and the vehicle is moving towards the radar set. An output from the comparator results only for the latter case.

Figure 3-16 shows the change in output voltage of A_9 versus time for a vehicle with constant speed. Mathematically, this can be shown by referring to Figure 3-12 and analyzing the equation

$$F_D = 2 F_o \frac{V_R}{C} . \quad (3-13)$$

From Figure 3-12

$$\theta = \tan^{-1} [Y/(X_o - X_1)] . \quad (3-14)$$

For

$$V = \text{Constant}$$

$$X_1 = V \int_{t_1}^{t_2} dt = V(t_2 - t_1) . \quad (3-15)$$

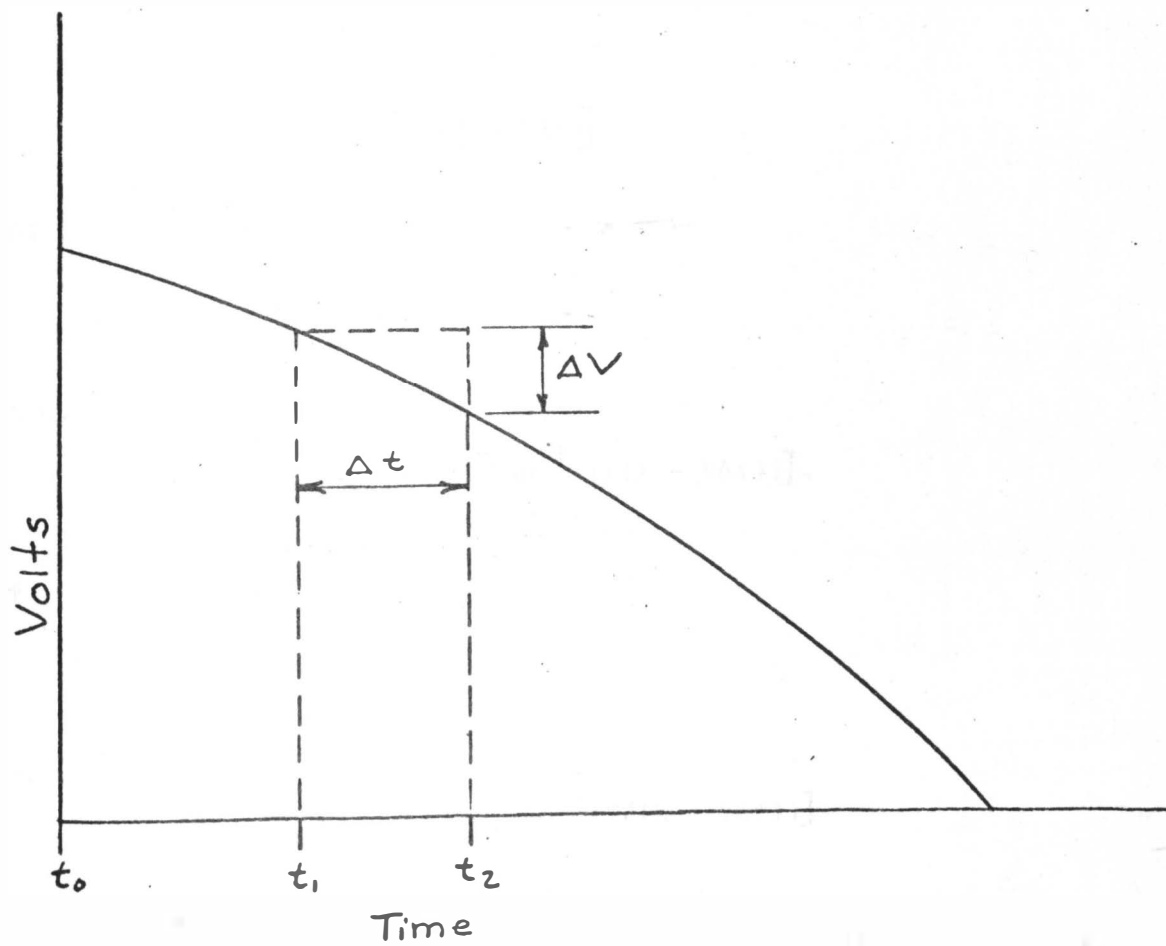


FIGURE 3-16. A_9 OUTPUT VOLTAGE VERSUS TIME FOR A CONSTANT VEHICLE SPEED.

Let,

$$\Delta t = t_1 - t_0,$$

$$\Delta t' = t_2 - t_1,$$

$$\theta = \tan^{-1} [Y/(X - V\Delta t)]$$

and

$$X_0 = X - V\Delta t.$$

Then,

$$F_D = \frac{2F_0 V}{C} \cos [\tan^{-1} (Y/(X - V\Delta t))].$$

If

$$\Delta F_D = F_D - F_D'$$

then

$$\begin{aligned} \Delta F_D &= \frac{2F_0 V}{C} \cos [\tan^{-1} (Y/(X - V\Delta t))] \\ &\quad - \frac{2F_0 V}{C} \cos [\tan^{-1} (Y/(X_0 - V\Delta t'))], \end{aligned}$$

rearranging

$$\begin{aligned} \Delta F_D &= \frac{2F_0 V}{C} \left[\cos [\tan^{-1} (Y/(X_0 - V\Delta t))] \right. \\ &\quad \left. - \cos [\tan^{-1} (Y/(X_0 - V\Delta t'))] \right] \end{aligned} \quad (3-16)$$

where

$$C = 1.86(10^5) \text{ miles/sec.} = 669.6(10^6 \text{ mph}),$$

$$Y = 50 \text{ ft.},$$

$$X_O = 500 \text{ ft.},$$

$$V = 30 \text{ mph} = 44 \text{ ft/sec.},$$

$$\Delta t' = 1 \text{ sec.},$$

and

$$F_O = 5(10^8) \text{ Hz.}$$

For $V = 30 \text{ mph}$, $X_O = 500 \text{ feet}$ and $\Delta t = 0$.

$$\begin{aligned} \Delta F_D &= \frac{2(5)10^8 (30 \text{ mph})}{669.6(10^6) \text{ mph}} \left[\cos \left[\tan^{-1} \left(\frac{50 \text{ feet}}{500 \text{ ft.} - 44 \text{ ft/sec}(0 \text{ sec})} \right) \right] \right. \\ &\quad \left. - \cos \left[\tan^{-1} \left(\frac{50 \text{ feet}}{500 \text{ ft.} - 44 \text{ ft/sec}(1 \text{ sec})} \right) \right] \right] \\ &= 44.8 [\cos(5.7106) - \cos(6.2574)] \\ &= 44.8 [.9950372 - .9940422] \\ &= 44.8 (9.95)10^{-4} \\ &= 0.045 \text{ Hz.} \end{aligned}$$

For $V = 30 \text{ mph}$ and $X_O = 200 \text{ feet}$

$$\begin{aligned} \Delta F_D &= 44.8 \left[\cos \left(\tan^{-1} \left(\frac{50}{200} \right) \right) - \cos \left(\tan^{-1} \left(\frac{50}{200 - 44} \right) \right) \right] \\ &= 44.8 [\cos(14.036) - \cos(16.763)] \\ &= 44.8 [.9701425 - .9575082] \\ &= 44.8 (.01263) \end{aligned}$$

$$= 0.566 \text{ Hz.}$$

For $V = 80 \text{ mph}$ and $X_O = 500 \text{ feet}$

$$\begin{aligned} \Delta F_D &= 44.8 \left[\cos \left[\tan^{-1} \left(\frac{50}{500} \right) \right] - \cos \left(\tan^{-1} \left(\frac{50}{500 - 117} \right) \right) \right] \\ &= 44.8 [.995037 - .991586] \\ &= 44.8 (3.45) 10^{-3} \\ &= 0.1546 \text{ Hz.} \end{aligned}$$

For $V = 80 \text{ mph}$ and $X_O = 300 \text{ feet}$

$$\begin{aligned} \Delta F_D &= 44.8 \left[\cos \left(\tan^{-1} \left(\frac{50}{300} \right) \right) - \cos \left(\tan^{-1} \left(\frac{50}{300 - 117} \right) \right) \right] \\ &= 44.8 [.99503 - .96464] \\ &= 44.8 (.0304) \\ &= 1.36 \text{ Hz.} \end{aligned}$$

The smallest change in voltage (ΔV) at the output of A_9 will be for a vehicle at $X_O = 500 \text{ feet}$ and $V = 30 \text{ mph}$. The slope of the curve in Figure 3-14 between 45 Hz and 94 Hz is

$$\begin{aligned} \text{Slope} &= \frac{-21.2V}{49 \text{ Hz}} \\ &= 0.433 \text{ V/Hz.} \end{aligned}$$

From previous calculations

$$\Delta F_D = 0.045 \text{ Hz,}$$

Therefore,

$$\begin{aligned}\Delta_v &= (-0.433 \text{ V/Hz})(0.045\text{Hz}) \\ &= 19.5 \text{ mv.}\end{aligned}$$

Using sufficient care to suppress the receiver noise level and minimize the input-offset voltage of A_9 , a voltage change this small can be detected by the voltage comparator A_{12} . The negative input to the comparator must be biased slightly above ground so for non-signal input the comparator output will be negative.

Examining Figure 3-14 it can be seen that for receding vehicles (F_D increasing) a positive voltage change causing a false indication will happen for two conditions. When the system locks onto a vehicle moving less than 47 mph (70 Hz) a positive voltage change is present at the input of the direction detector. Also, when a vehicle is moving faster than 47 mph a positive voltage change is made when the system falls out of lock.

A solution to this problem is presented in Figure 3-17. An output is desired only if the following three conditions are met.

1. A signal having a magnitude greater than some predetermined value is present at the input of the PLL.
2. The output of A_9 is increasing in a positive direction indicating F_D is decreasing.
3. The input signal has a frequency between 45 Hz and 94 Hz.

Condition 1 is met by sampling the input signal to the PLL, amplifying it and inputting it to an analog comparator through a diode

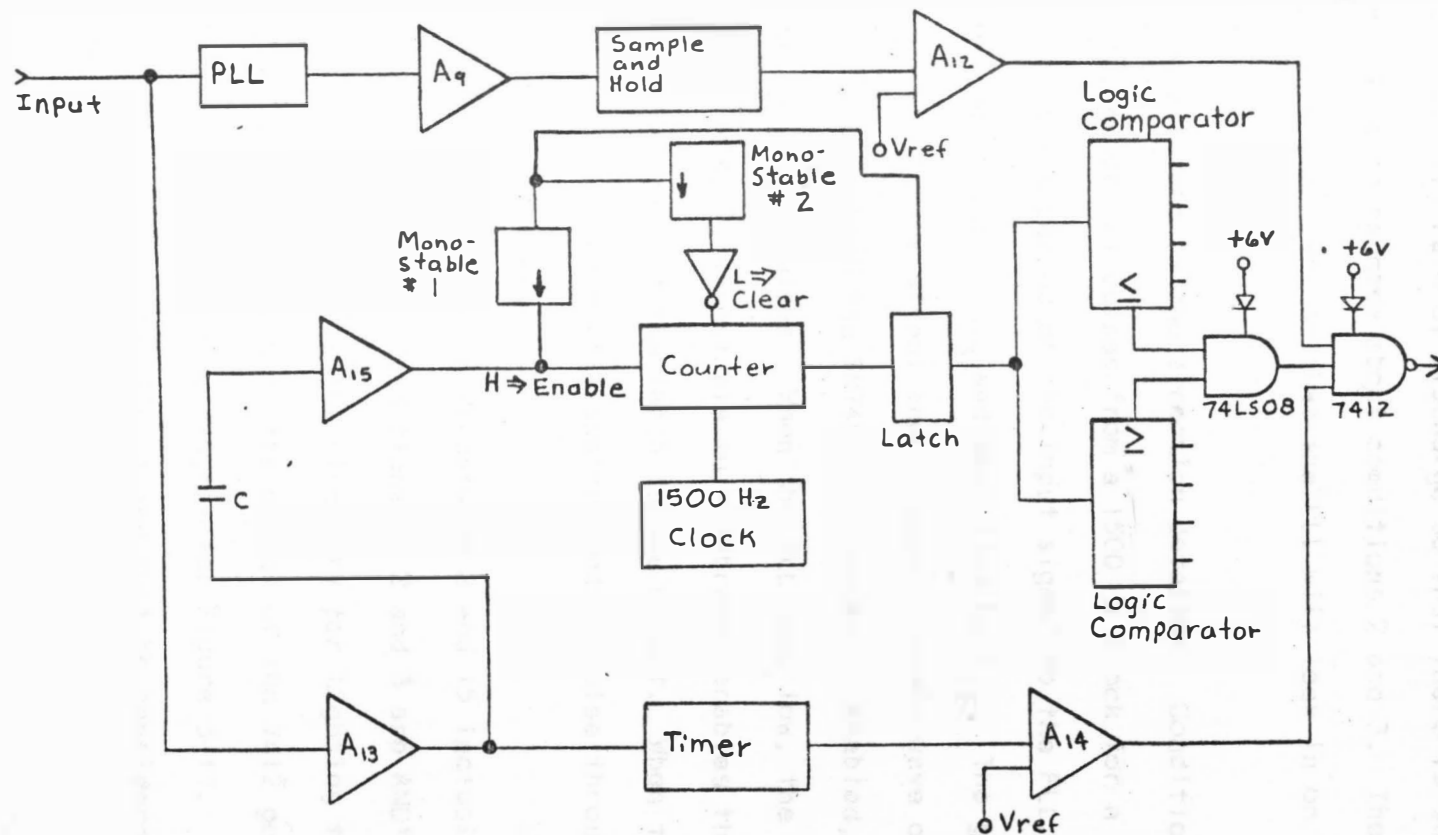


FIGURE 3-17. BLOCK DIAGRAM OF DIRECTION DETECTOR.

rectifier and capacitor filter in the timer. A resistor paralleling the capacitor controls its rate of discharge so that there is sufficient time for decisions to be made about conditions 2 and 3. The gain of A_{13} and V_{ref} of A_{14} are set so that the PLL will lock in on the signal before A_{14} will respond.

Condition 2 is met by the direction detector. Condition 3 is met by counting the number of pulses from a 1500 Hz clock for a time period equal to one half the period of the input signal to the PLL. The signal is taken from the output of A_{13} and amplified by A_{15} . The gain of A_{15} is sufficient to clip the signal and produce a square wave output. When this signal goes high the SN74LS196 counter is enabled, which then counts the NE 555 clock pulses. When the PLL goes low, the counter stops counting and an SN74L121 monostable multivibrator enables the SN74L75 latch shifting the input of the latch to the output. When the multivibrator goes low a second multivibrator sends a pulse through an inverter to clear the counter.

If the counter of the latch is between 8 and 15 inclusive the output of the 74L508 goes high. Conditions 1, 2 and 3 are AND'ed in an SN7412 NAND gate which controls the circuitry for lighting the indicator. If all conditions are high the output of the 7412 goes low.

Figure 3-18 is a timing diagram for Figure 3-17. Delay times are not shown but they are significant and must be considered.

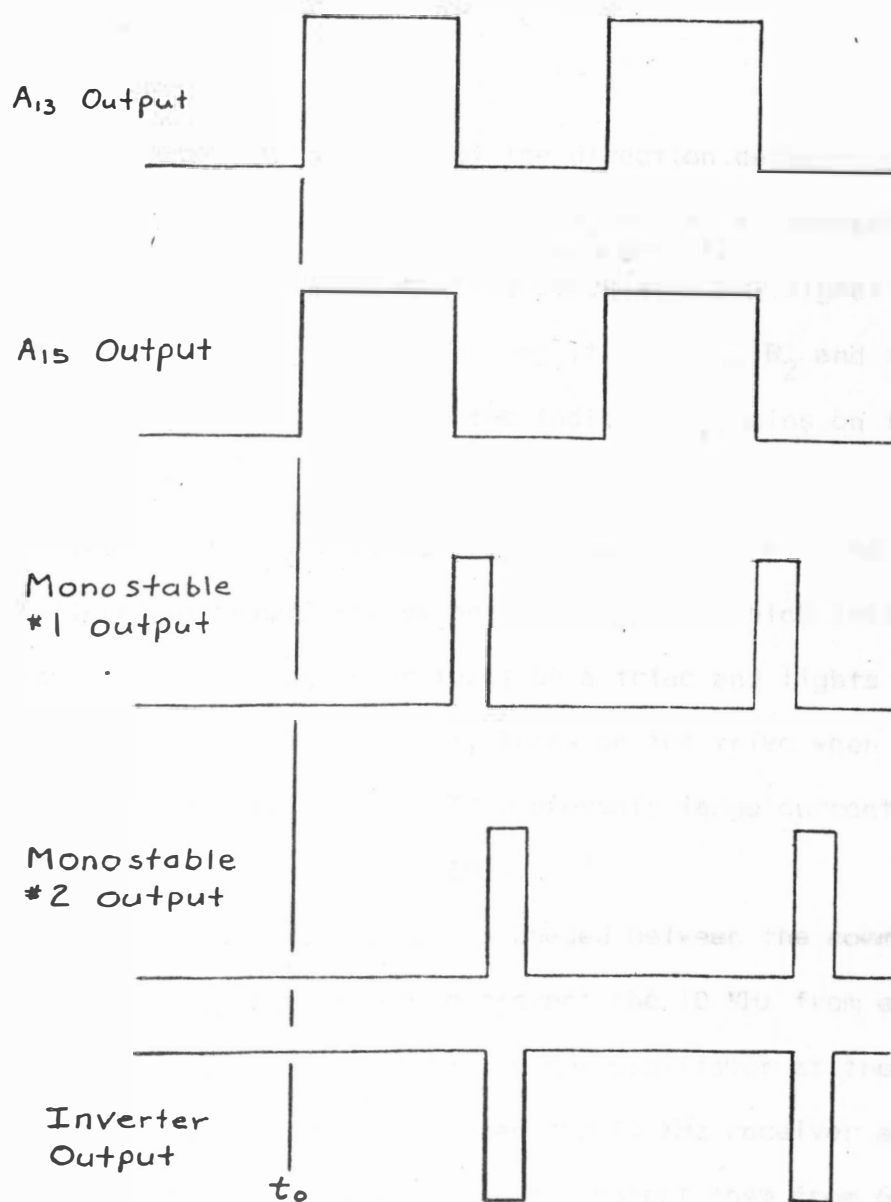


FIGURE 3-18. TIMING DIAGRAM FOR DIRECTION DETECTOR.

CHAPTER IV

INDICATOR

4-1. Indicator Operation.

When the output voltage of the direction detector goes low capacitor C_1 of Figure 4-1 discharges causing A_1 and A_2 to conduct. A_2 acts as a switch for the oscillator which sends a 10 KHz signal onto the a-c power line to the indicator turning it on. R_1 , R_2 and I_b of A_1 control the charge time of C_1 so that the indicator remains on for five seconds after the vehicle has passed the transmitter.

Figure 4-2 is the circuitry at the indicator. The 10 KHz signal is filtered and amplified by an active filter which activates a photo transistor. The transistor turns on a triac and lights the indicator. The zero crossing detector only turns on the triac when the 60 Hz voltage crosses the zero axis (21). This prevents large current surges which would cause noise radiation (EMI).

An isolation transformer is needed between the commercial line and the line feeding the system to prevent the 10 KHz from entering the main line. Filters between the 10 KHz oscillator at the transmitter and the a-c power line and between the 10 KHz receiver at the indicator and the a-c power line are needed to protect them from 60 Hz and line transients.

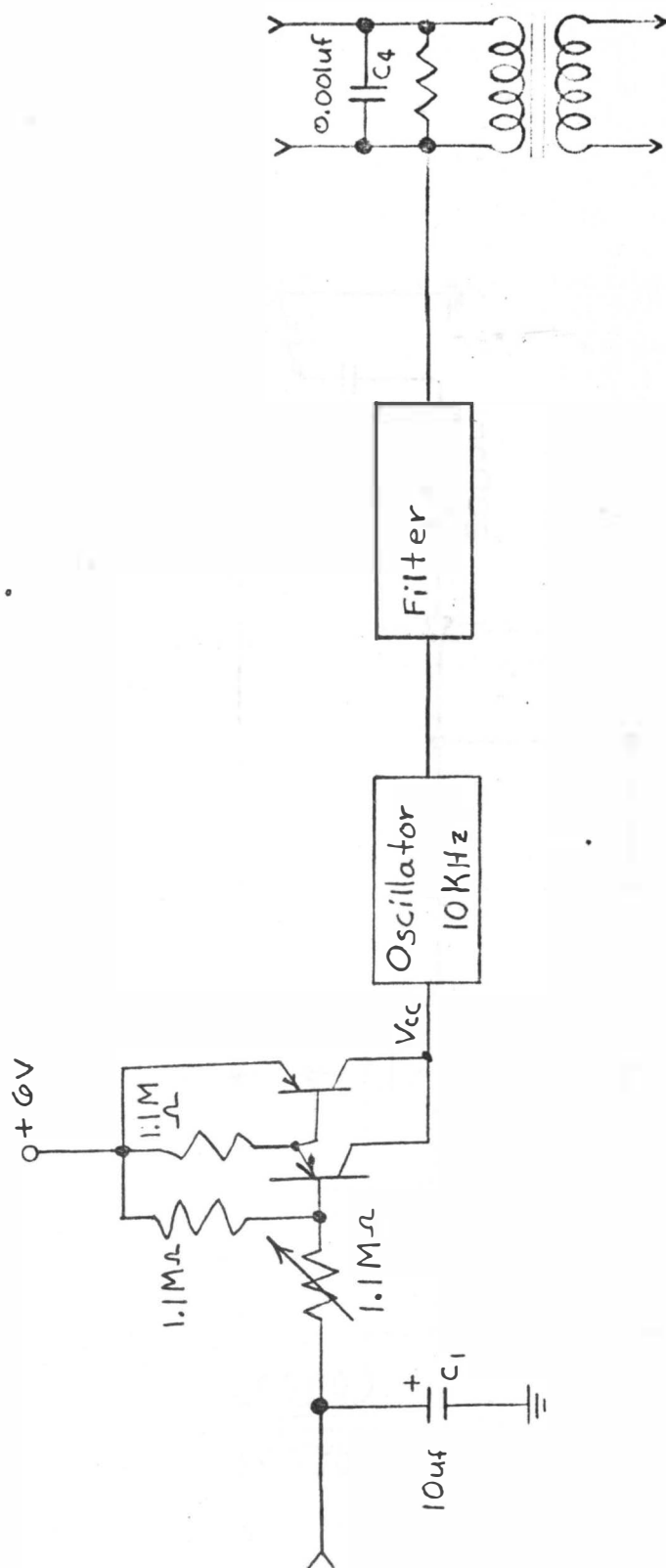


FIGURE 4-1. 10 KHz TRANSMITTER BLOCK DIAGRAM.

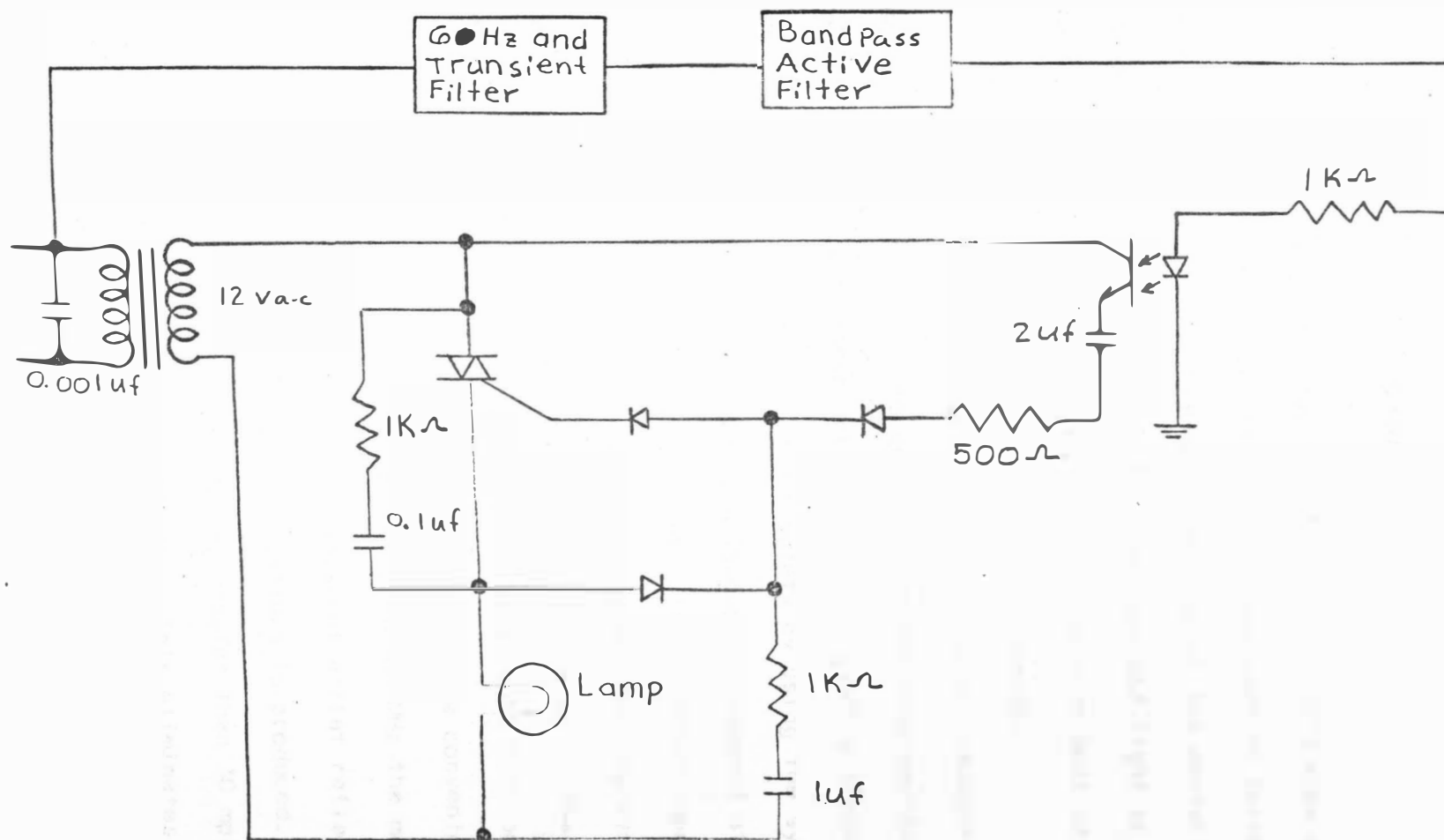


FIGURE 4-2. BLOCK DIAGRAM OF INDICATOR CIRCUITRY.

CHAPTER V

SUMMARY

The problem of traffic hazards obscuring a motorist's view at intersections does not have to be accepted as "too expensive to correct". A small doppler radar set can be placed on the side of the obstacle away from the intersection to detect on-coming vehicles and light an indicator located at the intersection. The set would have to meet state and Federal regulations and a license for its use is needed.

The use of this system will be primarily on rural roadways where high vehicle speeds, hills, curves and uncontrolled intersections exist. Private individuals having their driveways exit a short distance from the crest of a hill would increase their safety by using the system.

The designed system will operate continuously-on-demand at 500 MHz, 0.1 watt and have a range of about 500 feet. It is battery operated, +/- 12 volts and +/- 6 volts, with an a-c operated charging system. The transmitter and receiver electronics can be located in the same enclosure as the antenna, which can be a standard rural mail box or a metal enclosure having similar dimensions. It is more convenient to house the batteries in a separate container. At 500 MHz the mail box enclosure has characteristics similar to those of a flat reflector. Test results show that a good symmetrical pattern is produced.

Only vehicles having a radial velocity greater than 30 mph and approaching the transmitter will be detected. This eliminates any ambiguity that would otherwise exist.

The critical design aspects of the receiver were built and tested. The transmitter and power supply were designed. The indicator is a straight forward design using ordinary circuitry as found in handbooks. The tests on the receiver show that it performed as the design intended. The output of a sine wave generator was attenuated with a resistor divider network to $30\ \mu\text{v}$ and coupled into the receiver. The output of the sample-and-hold comparator was monitored with a voltmeter. Within the range from 45 Hz to 94 Hz decreasing frequencies were detected while increasing frequencies were not. Due to the limited range of the particular PLL used for the frequency to voltage converter, vehicles travelling faster than 63 mph, 8 mph over the speed limit, will not be detected as early as slower moving vehicles. Vehicles moving 80 mph will not be detected until they are 63 feet from the transmitter. These problems can be eliminated by placing the transmitter further from the indicator which will provide earlier detection.

If a complete system were constructed there would probably be a few refinements necessary in the design, however, the author is confident that the system will perform adequately as all critical circuitry has been designed, bread boarded and tested.

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